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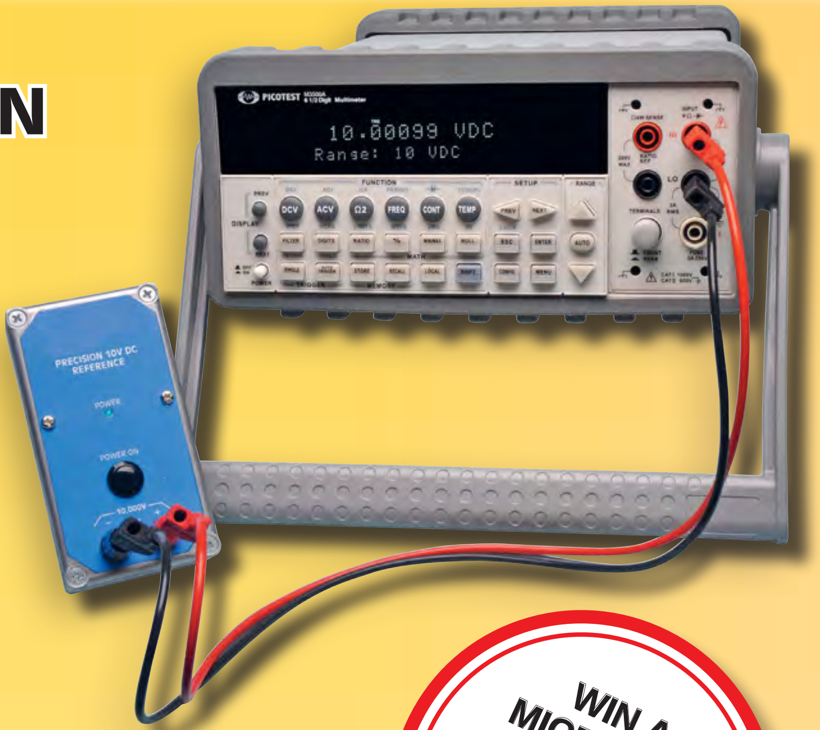

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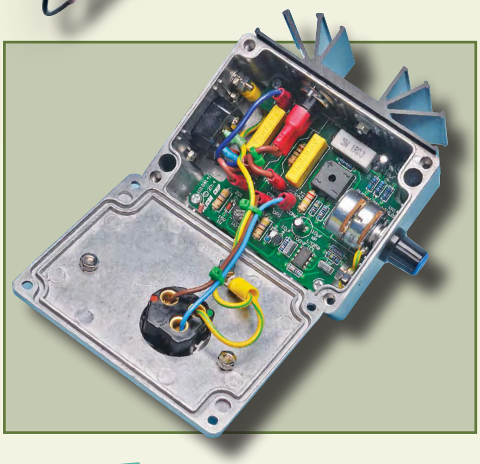

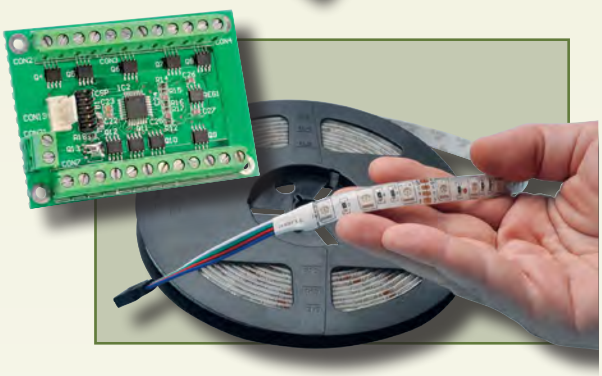

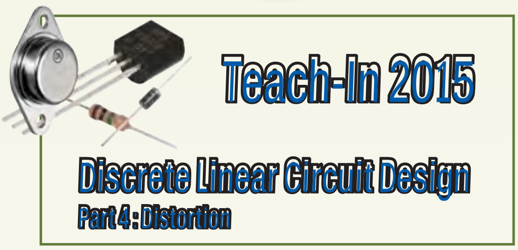

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*Assembled Order Code: VM111 - **£39.12***

Controllers & Loggers

Here are just a few of the controller and data acquisition and control units we have. See website for full details. 12Vdc PSU for all units: Order Code 660.446UK £11.52

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5 digital input channels and 8 digital output channels plus two analogue inputs and two analogue outputs with 8 bit resolution.
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State-of-the-art high security. 2 channel. Momentary or latching relay output rated to switch up to 240Vac @ 10 Amps. Range up to 40m. Up to 15 Tx's can be learnt by one Rx (kit includes one Tx but more available separately). 3 indicator LEDs. Rx: PCB 88x60mm, supply 9-15Vdc.
*Kit Order Code: 8157KT - **£49.95***
*Assembled Order Code: AS8157 - **£54.95***

Computer Temperature Data Logger

Serial port 4-channel temperature logger. °C or °F. Continuously logs up to 4 separate sensors located 200m+ from board. Wide range of free software applications for storing/using data. PCB just 45x45mm. Powered by PC. Includes one DS1820 sensor.
*Kit Order Code: 3145KT - **£19.95***
*Assembled Order Code: AS3145 - **£26.95***
*Additional DS1820 Sensors - **£4.95 each***

Remote Control Via GSM Mobile Phone

Place next to a mobile phone (not included). Allows toggle or auto-timer control of 3A mains rated output relay from any location

Most items are available in kit form (KT suffix) or pre-assembled and ready for use (AS prefix).

4-Ch DTMF Telephone Relay Switcher

Call your phone number using a DTMF phone from anywhere in the world and remotely turn on/off any of the 4 relays as desired. User settable Security Password, Anti-Tamper, **Rings** to Answer, Auto Hang-up and Lockout. Includes plastic case. 130 x 110 x 30mm. Power: 12Vdc.
*Kit Order Code: 3140KT - **£79.95***
*Assembled Order Code: AS3140 - **£94.95***

8-Ch Serial Port Isolated I/O Relay Module

Computer controlled 8 channel relay board. 5A mains rated relay outputs and 4 opto-isolated digital inputs (for monitoring switch states, etc). Useful in a variety of control and sensing applications. Programmed via serial port (use our new Windows interface, terminal emulator or batch files). Serial cable can be up to 35m long. Includes plastic case 130x100x30mm. Power: 12Vdc/500mA.
*Kit Order Code: 3108KT - **£74.95***
*Assembled Order Code: AS3108 - **£89.95***

Infrared RC 12–Channel Relay Board

Control 12 onboard relays with included infrared remote control unit. Toggle or momentary. 15m+ range. 112 x 122mm. Supply: 12Vdc/0.5A
*Kit Order Code: 3142KT - **£64.95***
*Assembled Order Code: AS3142 - **£74.95***

Audio DTMF Decoder and Display

Detect DTMF tones from tape recorders, receivers, two-way radios, etc using the built-in mic *or* direct from the phone line. Characters are displayed on a 16 character display as they are received and up to 32 numbers can be displayed by scrolling the display. All data written to the LCD is also sent to a serial output for connection to a computer. Supply: 9-12V DC (Order Code PSU375). Main PCB: 55x95mm.
*Kit Order Code: 3153KT - **£37.95***
*Assembled Order Code: AS3153 - **£49.95***

3x5Amp RGB LED Controller with RS232

3 independent high power channels. Preprogrammed or user-editable light sequences. Standalone option and 2-wire serial interface for microcontroller or PC communication with simple command set. Suitable for common anode RGB LED strips, LEDs and incandescent bulbs. 56 x 39 x 20mm. 12A total max. Supply: 12Vdc.
Kit Order Code: 8191KT - £29.95
Assembled Order Code: AS8191 - £39.95

Hot New Products!

Here are a few of the most recent products added to our range. See website or join our email Newsletter for all the latest news.

4-Channel Serial Port Temperature Monitor & Controller Relay Board

4 channel computer serial port temperature monitor and relay controller. Four inputs for Dallas DS18S20 or DS18B20 digital thermometer sensors (£3.95 each). Four 5A rated relay outputs are independent of sensor channels allowing flexibility to setup the linkage in any way you choose. Simple text string commands for reading temperature and relay control via RS232 using a comms program like Windows HyperTerminal or our free Windows application.
Kit Order Code: 3190KT - ***£84.95***
Assembled Order Code: AS3190 - ***£99.95***

40 Second Message Recorder

Feature packed non-volatile 40 second multi-message sound recorder module using a high quality Winbond sound recorder IC. Standalone operation using just six onboard buttons or use onboard SPI interface. Record using built-in microphone or external line in. 8-24Vdc powered. Change a resistor for different recording duration/sound quality. Sampling frequency 4-12 kHz. (120 second version also available)
Kit Order Code: 3188KT - ***£29.95***
Assembled Order Code: AS3188 - ***£37.95***

Bipolar Stepper Motor Chopper Driver

Get better performance from your stepper motors with this dual full bridge motor driver based on SGS Thompson chips L297 & L298. Motor current for each phase set using on-board potentiometer. Rated to handle motor winding currents up to 2 Amps per phase. Operates on 9-36Vdc supply voltage. Provides all basic motor controls including full or half stepping of bipolar steppers and direction control. Allows multiple driver synchronisation. Perfect for desktop CNC applications.
Kit Order Code: 3187KT - ***£39.95***
Assembled Order Code: AS3187 - ***£49.95***

Video Signal Cleaner

Digitally cleans the video signal and removes unwanted distortion in video signal. In addition it stabilises picture quality and luminance fluctuations. You will also benefit from improved picture quality on LCD monitors or projectors.
Kit Order Code: K8036 - ***£24.70***
Assembled Order Code: VM106 - ***£36.53***

Motor Speed Controllers

Here are just a few of our controller and driver modules for AC, DC, Unipolar/Bipolar stepper motors and servo motors. See website for full details.

DC Motor Speed Controller (100V/7.5A)

Control the speed of almost any common DC motor rated up to 100V/7.5A. Pulse width modulation output for maximum motor torque at all speeds. Supply: 5-15Vdc. Box supplied. Dimensions (mm): 60Wx100Lx60H.
Kit Order Code: 3067KT - ***£19.95***
Assembled Order Code: AS3067 - ***£27.95***

Bidirectional DC Motor Speed Controller

Control the speed of most common DC motors (rated up to 32Vdc/10A) in both the forward and reverse direction. The range of control is from fully OFF to fully ON in both directions. The direction and speed are controlled using a single potentiometer. Screw terminal block for connections.
Kit Order Code: 3166v2KT - ***£23.95***
Assembled Order Code: AS3166v2 - ***£33.95***

Computer Controlled / Standalone Unipolar Stepper Motor Driver

Drives any 5-35Vdc 5, 6 or 8-lead unipolar stepper motor rated up to 6 Amps. Provides speed and direction control. Operates in stand-alone or PC-controlled mode for CNC use. Connect up to six 3179 driver boards to a single parallel port. Board supply: 9Vdc. PCB: 80x50mm.
Kit Order Code: 3179KT - ***£17.95***
Assembled Order Code: AS3179 - ***£24.95***

Computer Controlled Bi-Polar Stepper Motor Driver

Drive any 5-50Vdc, 5 Amp bi-polar stepper motor using externally supplied 5V levels for STEP and DIRECTION control. Opto-isolated inputs make it ideal for CNC applications using a PC running suitable software. Board supply: 8-30Vdc. PCB: 75x85mm.
Kit Order Code: 3158KT - ***£24.95***
Assembled Order Code: AS3158 - ***£34.95***

AC Motor Speed Controller (600W)

Reliable and simple to install project that allows you to adjust the speed of an electric drill or 230V AC single phase induction motor rated up to 600 Watts. Simply turn the potentiometer to adjust the motors RPM. PCB: 48x65mm. Not suitable for use with brushless AC motors.
Kit Order Code: 1074KT - ***£15.95***
Assembled Order Code: AS1074 - ***£23.95***

See website for lots more DC, AC and stepper motor drivers!

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40 MHz sampling rate, 12 MHz analog bandwith, 0.1 mV sensitivity, 5mV to 20V/div in 12 steps, 50ns to 1 hour/div time base in 34 steps, ultra fast full auto set up option, adjustable trigger level, X and Y position signal shift, DVM readout and more...
Order Code: HPS50 - ~~£289.96~~ ***£203.95***

See website for more super deals!

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EVERYDAY PRACTICAL ELECTRONICS

10V, precisely

When it comes to measurement, there is nothing inherently 'fundamental' about voltage; other electronic parameters such as current, charge or frequency can and are used to represent variables and physical quantities. However, voltage measurement really is the bread and butter of electronic instrumentation, and it is only appropriate that we acknowledge this with a project that lets you calibrate your meters and circuit references with a master reference. Our *Low-cost Precision Reference* – without any adjustment – will provide you with a source of 10.000V DC, accurate to within ±5mV or ±0.05%. Just what you need to check your meters are within spec and your designs are performing to their best.

Of course, this month's other projects are equally fascinating. The *Deluxe Fan Speed Controller*, as the description points out, is equally at home acting as a lamp dimmer and I am sure you could find many other ingenious uses for this compact noise-free circuit.

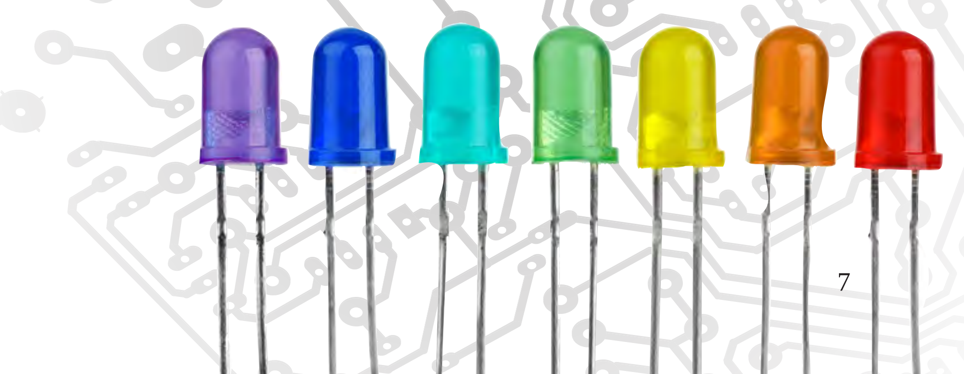
The *RGB LED Strip Driver* is a nice way to take advantage of flexible LED 'ribbons' that enable you to produce a 'rainbow' of colours. It's the kind of project where there are so many options you are only limited by your own imagination!

Ge-mania

I hope you are enjoying Jake Rothman's series on germanium technology. Jake tells me that one of the main motivations for the series was a lack of readily-available information on germanium devices, which he uses in some of the original audio products he designs and manufactures. I very much doubt you'll find a richer vein of handy tips and tricks on using this, the 'granddaddy material' of solid-state electronics, anywhere else than in *EPE*. It's fascinating to see how designers used early transistors in audio applications, and as always, Jake provides his own unique, hands-on approach to describing technology.

Something for nothing?

Well, almost! Last, but not least, a quick reminder to all *EPE* readers that for just a minute of your time you can enter our online competition, generously supported by Microchip. Each month, the free-to-enter competition offers you a chance to win some excellent prizes – see page 19 for May's exciting prize and details of entry.



NEWS

A roundup of the latest Everyday News from the world of electronics

TV's future starts to take shape... report by Barry Fox

Korean companies Samsung and LG now dominate the TV market. Previous leaders such as Philips, Sharp, Toshiba, Mitsubishi and Hitachi have either stopped making sets or sold out to Chinese cashlords. Sony is in corporate turmoil. Panasonic remains – for the time being at least – an innovative player, and is heavily betting on the success of 4k Ultra High Definition. But, as the annual Panasonic European Convention, held this year in Frankfurt, showed, the lack of a future-proof standard for UHD is crippling the industry.

HDR issues

During the plenary session much was made of the 'need for better pixels, not just more pixels', with High Dynamic Range generally regarded as the best bet for 'bettering' because its effect is immediately eye-catching in a shop. Panasonic was showing a prototype TV with a high brightness (1000 nit or candela per square metre) screen capable of delivering HDR with very wide graded contrast between extreme black and white. But several industry groups are still studying several different HDR control systems, to try and agree a standard for HDR. So there is no way of knowing whether high brightness sets sold now will meet whatever HDR standard is set.

The message to press and dealers attending the Convention was very muddled, too. One demonstrator assured that 'the HDR standards have now been set – a factory engineer told me three days ago'. Another admitted he was not sure about standards and did an Internet search with his tablet, which confirmed that there were still several proposals on the table, and no immediate prospect for agreement.

To add to the overall confusion over HDR, the prototype 1000 nit screen was in a darkened room – rather than real world ambience – with next to no technical information and it was displaying pictures that were sometimes so over-bright that viewers were wincing and turning away.

Last year, Panasonic was first to adopt and adapt the Freetime satellite PVR system to provide a backwards-and-forwards-looking EPG for 4k TVs. But Panasonic is now adopting the recently announced Freeview Play system, which is claimed to offer a more appealing user look-and-feel. Freeview Play is expected to be ready for implementation by May and Panasonic pledges that any sets sold ahead of system finalisation will be software-upgradable.

Operating systems

Panasonic also announced that it will use the Mozilla Firefox Open Operating System for new smart/connected TVs. This follows the decision by rival Sony to use the Android OS, and is in keeping with the traditional we-do-it-differently competition between the two companies.

Live demonstrations showed how, when the Firefox TV set is first switched on, it displays three circular coloured icon Widgets, for Live TV, Apps and Devices. The owner can then 'pin' similar icons to the Home screen, which provide short cuts to favourite TV channels, connected devices such as a BD player or catch-up TV services such as BBC iPlayer.

A panel session with Paul Gray, director of European research for DisplaySearch/IHS and Andreas Gal, Mozilla's chief technical officer, gave some insight into why TV manufacturers are now turning to computer operating systems.

Development strategy

Said Paul Gray: 'Developing software is like a tax. The makers face a nasty dilemma. You have to keep developing to be successful but if you do it yourself you spend a huge amount of money. Then you make a loss. So it makes sense not to do it all on your own. There are people out there who have huge amounts of knowledge, for instance on security. It makes sense to involve them. This is a pragmatic move forward.

'The early TVs just had a tablet inside them. A lot of the software was location specific. But a TV is not a 55-inch smartphone. It's not a mobile touch screen. People are still working out the right user experience. There isn't an accepted right answer yet.

'TVs have always used Open Source Linux so this adoption of open source Firefox is just a continuation of that path. A lot of the infrastructure work has been done. This is the next stage.

'The whole point of connected devices is that they talk to each other. If we continue with silos and stacks not talking to each other, most consumers won't bother. Consumers want a frictionless experience. Everything must talk to everything else. Open source can make this happen."

Said Andreas Gal: 'Panasonic does not pay us. It's completely open source. If YOU want to start making TVs with Firefox you can download it from our website.

'But don't make the mistake of thinking that because Mozilla is a non-profit organisation, no-one is making money. We very much want Panasonic and other OEMs to make money because businesses need to make money. In the smartphone space we already understand pretty well how to monetise. You drive a car, I can monetise that. You are hungry, I can monetise that. In the smart TV space it's less clear. But with connected TVs people will consume more interactive services and many of those are monetisable. So we are focussed on ensuring there is a healthy ecosystem.'

Discovery of rare part of early Cambridge computer

A very rare original part of EDSAC, one of the world's first computers, has resurfaced in the US and the discovery suggests that other parts of EDSAC may still be in existence.

The part, a chassis designed to hold 28 of the 3000 EDSAC valves, has just been donated to the EDSAC team at The National Museum of Computing (TNMOC), where the ongoing reconstruction of EDSAC, originally built in Cambridge in the late 1940s, is on display. The reconstruction is expected to be completed later this year and is already a very popular exhibit, especially among the many educational groups that visit the Museum.

The EDSAC Chassis part is said to have been acquired at some sort of auction of EDSAC parts in Cambridge in the 1950s when the computer was decommissioned.

Andrew Herbert, the leader of the EDSAC project at TNMOC, said: 'Details of the "auction" are unclear, but there is a possibility that other parts of the original EDSAC still exist and could even be in the Cambridge area stored away in lofts, garden sheds and garages. We would very much like to hear from anyone who thinks they may have other parts.'

EDSAC, the Electronic Delay Storage Automatic Calculator, was built immediately after World War II by a team led by Sir Maurice Wilkes in the Mathematical Laboratory of the University of Cambridge. It was one of the first practical general purpose computers and was used by scientific researchers across the University. The EDSAC design was later developed to create LEO, the world's first business computer.

You can follow progress of the reconstructuction of EDSAC at: **www.tnmoc.org/special-projects/edsac**

Handy Zener tester

Peak Electronic Design Ltd has announced its latest product, the Atlas ZEN. This new instrument allows for the rapid and accurate testing of Zener diodes (including Avalanche diodes) up to 50V. The instrument will even measure the slope resistance of Zeners under test.

Test currents are selectable from 2mA, 5mA, 10mA and 15mA (which suits most Zeners). The duty cycle of the test current is kept very low, so there is negligible heating of the diode under test, even for a 50V Zener at 15mA.

It is unusual for any tester to be able to measure Zener voltages up to 50V and it is even more unusual for any tester to be able to simultaneously measure slope resistance. This can be particularly useful if you want to choose the optimum drive current for your Zener. The Atlas ZEN cleverly measures slope resistance by mathematically fitting a curve to the component's I/V characteristics, giving more accurate results than a simple 2-point straight line approximation.

Not only will the Atlas ZEN test Zeners, but it's also handy for other components too, such as LEDs (and LED strings up to 50V), transient voltage suppressors, shunt regulators and conventional diodes. Be careful not to test LEDs in reverse though as the LED won't cope with the reverse voltages generated by the Atlas ZEN!

The unit is powered by a single Alkaline AAA cell (included) and test conditions are regulated over the full battery life (down to 0.9V).

The Atlas ZEN is priced at just £39.00 (incl VAT), with UK shipping at £3. Orders can be placed at Peak's website: **www.peakelec.co.uk**

Beeb's Micro Bit

A major BBC project, developed in pioneering partnership with over 25 organisations, will give a personal coding device free to every child in year 7 across the country this autumn – one million devices in total.

Still in development and nicknamed the 'Micro Bit', the project builds on the legacy of the BBC Micro, which was put into schools in the 1980s and was instrumental in the careers of many of today's technology pioneers. Computing and digital technology has become ubiquitous since then, but for many, the emphasis has shifted from creation to consumption.

The Micro Bit will be a wearable device with an LED display that can be programmed. It will be a standalone, entry-level coding device that allows children to pick it up, plug it into a computer and start creating with it immediately. It will connect and communicate with other devices, including Arduino, Galileo, Kano and Raspberry Pi, as well as other Micro Bits.

Insect search and rescue

Cyborg insect researchers at UC Berkeley and Singapore's Nanyang Technological University (NTU) are strapping tiny computers and wireless radios onto the backs of giant flower beetles and recording neuromuscular data as the bugs fly untethered. The researchers then used that information to improve the precision of the beetles' remote-controlled turns.

Research in this field could lead to applications such as controlling the flight direction of sensor and camera-carrying insects to aid search-and-rescue operations in areas too dangerous for humans.

Low-cost precision 10V DC reference for checking DMMs

Ever checked the calibration of your digital multimeter? OK, we know you haven't because there's no easy or cheap way of doing it. But now you can, with this low-cost precision DC voltage reference. Without any adjustment it will provide you with a source of 10.000V DC, accurate to within ±5mV or ±0.05%.

By JIM ROWE

MOST OF US DON'T ever get our DMMs calibrated, though we know that they do drift out of calibration over years of use. However, if you are using them during the course of your work, they should be checked every year or so – otherwise how can you trust the readings?

The problem is, it can cost quite a lot to send a DMM away to a standards lab for calibration – more than many DMMs are worth. So generally we either hope for the best or simply buy a new DMM if we suspect that our existing meter has drifted too far out of calibration.

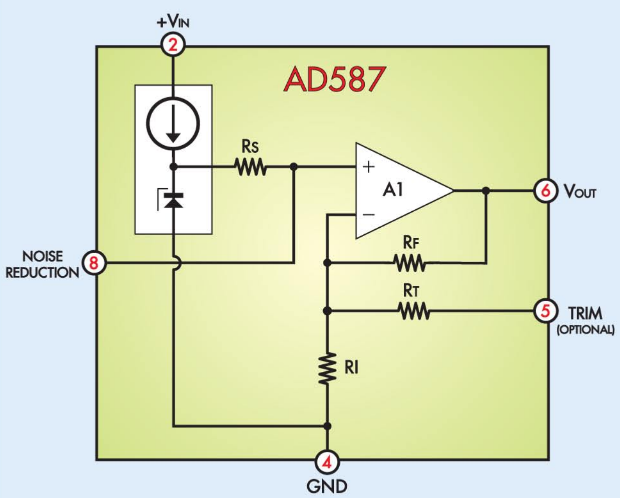
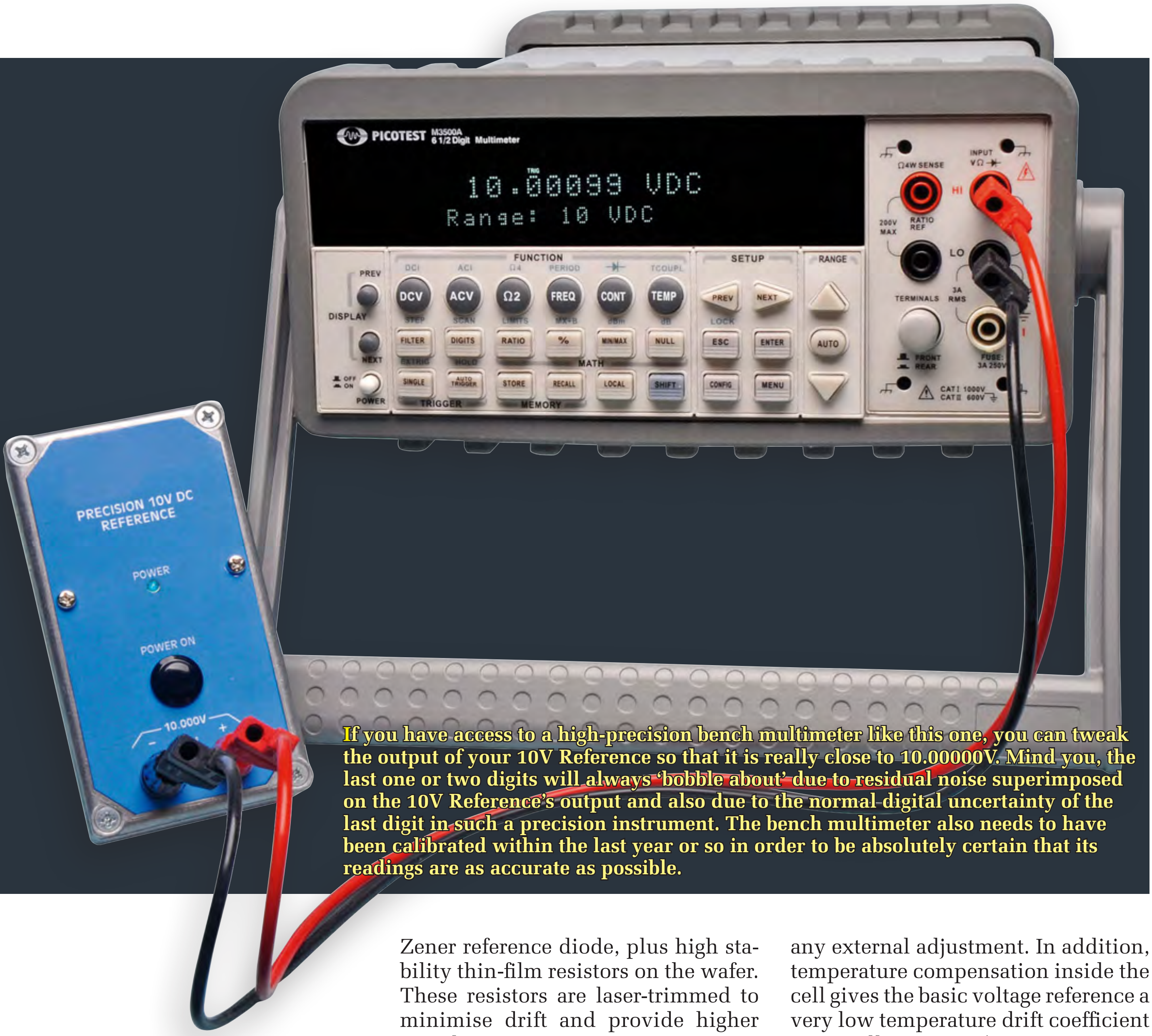


Fig.1: block diagram of the AD587 10V voltage reference. It consists of a buried zener diode and its associated current source, plus op amp IC1 which operates as an adjustable-gain buffer stage. Buried zeners have their avalanche zone several µm inside the oxide layer and so do not suffer from long-term drift or 'walkout'.

Back in the 1970s, when DMMs first became available, the only practical DC voltage reference was still the Weston cell. This wet chemical 'primary cell' had been developed in 1893 and subsequently became the international standard for EMF/voltage in 1911. It produced an accurate 1.0183V reference, which could be used to calibrate DMMs and other instruments.

Unfortunately, Weston cells were fairly expensive and few technicians had direct access to one for meter calibration. As a result, a reasonably-fresh mercury cell was often used as a kind of 'poor man's' voltage reference. Fresh mercury cells have a terminal voltage very close to 1.3566V at 20°C and the voltage falls quite slowly to about 1.3524V after a year or so. Silver oxide cells were also used for the same purpose, having a stable terminal voltage very close to 1.55V.

Of course, batteries have a tendency to obey 'Murphy's Law' and usually turn out to have quietly expired just before you need them. And although



If you have access to a high-precision bench multimeter like this one, you can tweak the output of your 10V Reference so that it is really close to 10.00000V. Mind you, the last one or two digits will always 'bobble about' due to residual noise superimposed on the 10V Reference's output and also due to the normal digital uncertainty of the last digit in such a precision instrument. The bench multimeter also needs to have been calibrated within the last year or so in order to be absolutely certain that its readings are as accurate as possible.

mercury and silver oxide cells have quite a long life, especially if you use them purely as a voltage reference, they certainly aren't immune to this problem. So these batteries make a pretty flaky voltage reference, at best.

Fortunately, in the 1980s, semiconductor makers developed a relatively low-cost source of stable and accurate DC voltage: the monolithic voltage reference (MVR). This is basically a very accurate voltage regulator. It produces a precise regulated DC output voltage when fed with unregulated DC power, but unlike the more familiar 3-terminal regulators, it can supply very little current.

Analog Devices' AD587, used in this new *Precision 10V Reference Mk.2*, incorporates a number of recent advances in MVR technology. These include an ion-implanted 'buried' Zener reference diode, plus high stability thin-film resistors on the wafer. These resistors are laser-trimmed to minimise drift and provide higher initial accuracy.

The AD587 also operates from an unregulated input voltage of between +15V and +18V, with a quiescent current of just 4mA. This is somewhat lower than earlier MVRs, making it very suitable for battery-powered operation.

Block diagram

Fig.1 shows what's inside an AD587. The voltage reference cell itself is at upper left, consisting of the 'buried' Zener and its current source.

The other main circuit section is op amp A1, used as an adjustable gain buffer. R_F, R_I and R_T are high-stability thin-film resistors, laser trimmed to allow the gain of A1 to be set with a high degree of precision. The output voltage (between pins V_{OUT} and GND) is initially set to 10.000V ±5mV for the AD587KNZ version used here, without any external adjustment. In addition, temperature compensation inside the cell gives the basic voltage reference a very low temperature drift coefficient – typically ±10ppm/°C.

Note that a slightly lower-spec version of the AD587 is also available, the AD587JNZ. This offers an initial (untrimmed) DC output voltage of 10.000V ±10mV, with a temperature drift coefficient of ±20ppm/°C. So you could use it as an 'almost as good' alternative if the KNZ version becomes unavailable.

Although the 'untrimmed' initial accuracy of the AD587KNZ (10.0V ±0.05%) is good enough for calibrating most low-cost DMMs, the chip can also be easily trimmed to improve its accuracy by a factor of greater than 10 times, ie, to around ±0.002%. This is done by connecting its TRIM pin (pin 5) to a trimpot circuit, connected between the V_{OUT} and GND terminals. This allows the gain of A1 to be adjusted to give an output anywhere within the range 9.900V to 10.300V,

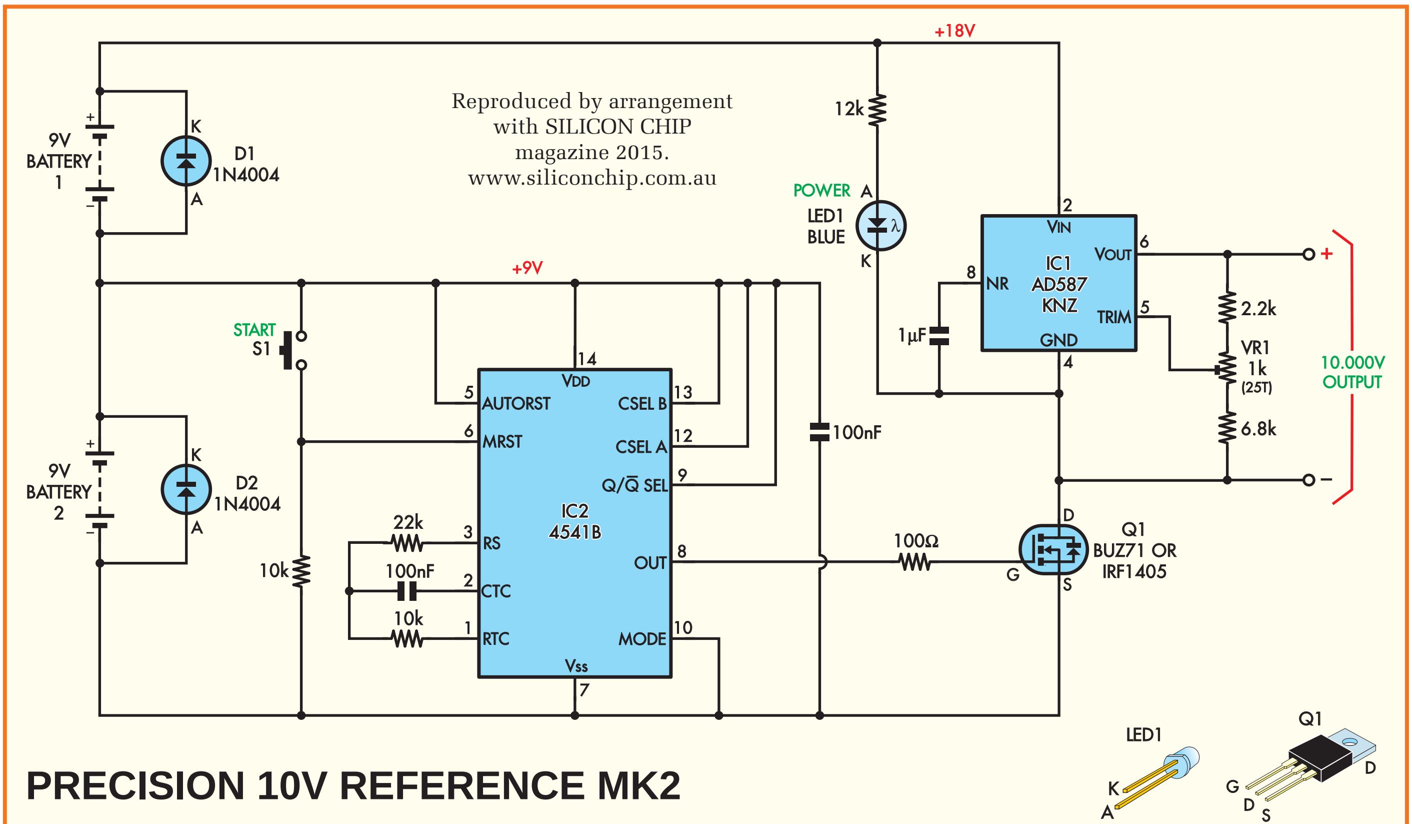


Fig.2: the complete circuit diagram. IC1 is the precision 10V reference, while IC2 operates as a 90s timeout counter. When S1 is pressed, IC2 turns MOSFET Q1 on for 90s and connects IC1 and LED1 across the 18V supply.

with no adverse effect on temperature stability.

If this trim adjustment range seems a little wide, this has been done deliberately to provide the option of setting the output voltage to 10.240V. It can then be used as a reference source for binary DACs and ADCs (more about this later).

The 400mV adjustment range does mean that in order to accurately set the output voltage, we have to use a 25-turn trimpot in series with two fixed resistors. And of course, in order to take advantage of this trimming feature, you really need access to an even higher precision voltage reference to compare it with. Either that, or access to a recently calibrated high-resolution DMM.

Circuit details

Refer now to Fig.2 for the complete circuit details. There's not a lot to it – just the AD587KNZ precision voltage reference (IC1) plus some extra circuitry to allow the AD587KNZ to run from two 9V alkaline batteries to provide a truly portable reference.

This additional circuitry is based around IC2, a programmable CMOS timer. It provides a 90-second timeout function and controls IC1's operation via Q1, a BUZ71 (or IRF1405) N-channel MOSFET.

IC2 (4541B) is basically a binary counter with 16 stages. It can be configured as either an 8, 10, 13 or 16-stage counter by changing the logic levels to which its two 'CSEL' programming inputs (pins 12 and 13) are connected. In this circuit, both these inputs have been connected to +9V (ie, tied high), to configure the counter to use its full 16 stages.

The 4541B also contains its own clock oscillator, the frequency of which is set by the RC timing components connected to pins 1, 2 and 3. In this case, the values specified give an overall timer period of around 85-90 secs.

IC2's output at pin 8 drives MOSFET Q1's gate via a 100Ω resistor. As a result, each time pushbutton switch S1 is pressed (and resets the counter), pin 8 of IC2 goes high and Q1 turns on and connects IC1 across the 18V supply for the duration of the 85-90s timing period. At the end of this period, pin 8 switches low and Q1 turns off to remove power from IC1 and conserve battery life.

Pressing S1 again starts the timing period all over again, if further calibration checks are necessary.

Power comes from the two 9V batteries, while D1 and D2 act as voltage clamps to provide reverse polarity protection if a battery is connected the wrong way around. LED1 and its associated 12kΩ current-limiting resistor are connected across IC1's supply pins, so the LED functions as a power-on indicator. Using a high-efficiency 3mm blue LED gives a very visible indication, while adding less than 1.5mA to the total current drain.

By the way, you may be wondering why we have used a BUZ71 or IRF1405 power MOSFET for Q1 when IC1 and LED1 only draw a maximum of 16mA or so, even with a 10mA external load (the maximum current the AD587 can provide). This is because the BUZ71 (or IRF1405) offers a much lower on-resistance than smaller low-power MOSFETs like the 2N7000. This provides a much lower voltage drop and allows us to achieve significantly longer life from the 9V batteries.

The connections for IC1 itself are easy to follow. The 1µF capacitor connected between pin 8 (NR) and pin 4

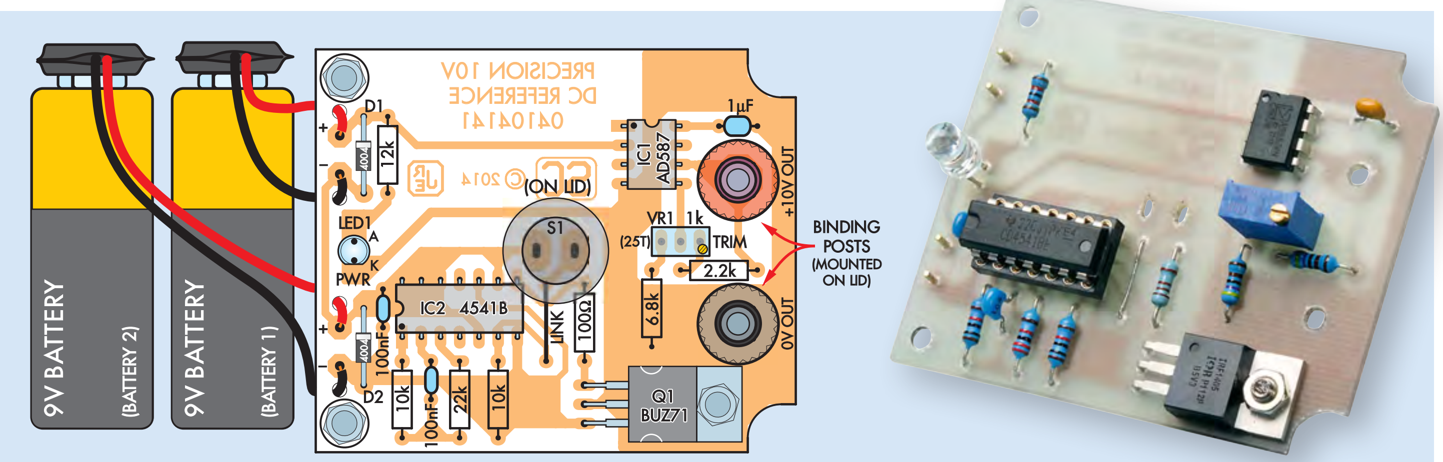


Fig.3: follow this layout diagram to build the unit, but note that switch S1 and the two binding post terminals are soldered to the PCB only after they have been mounted on the case lid (see text). Leave out trimpot VR1 and the 2.2kΩ and 6.8kΩ resistors if you don't intend calibrating the unit. Note: the prototype PCB shown in the photo lacks the reverse-polarity protection diodes and the strain relief holes for the battery leads included in the final version.

(GND) is there to provide additional low-pass filtering of any noise generated by the AD587's buried Zener. It works in conjunction with series resistor R_S, which is shown in Fig.1.

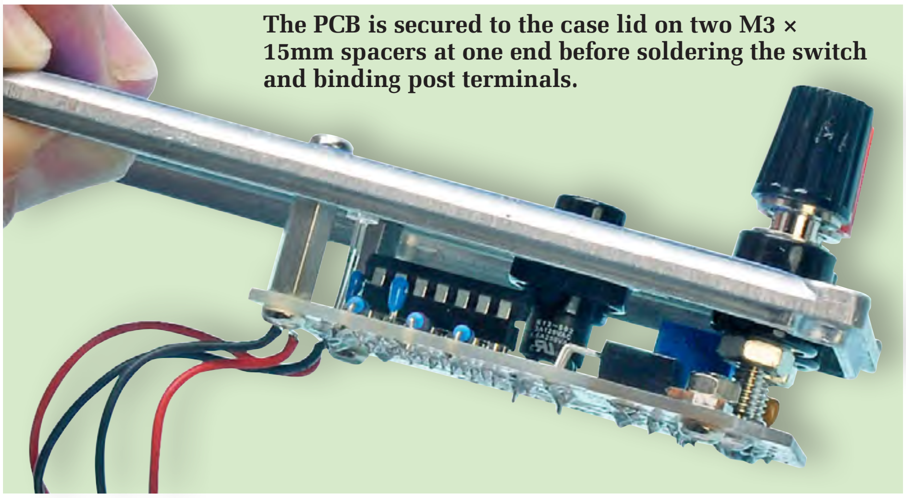
Trimpot VR1 and its two range-setting resistors are for 'trimming' the output voltage of IC1 to the desired 10.000V or 10.240V. However, note that there's no point in fitting these parts unless you have access to a very accurately-calibrated DMM, to compare it against while you're doing the trimming adjustment. **In fact, these parts must be left out if you have no way of performing the calibration, otherwise they will upset the accuracy.**

Conversely, if you are able to carry out calibration, the resistor values shown (2.2kΩ and 6.8kΩ) will give a trimming range centred on 10.000V. Alternatively, if you want the trimming range to be centred on 10.240V, change the 2.2kΩ 'upper' resistor to 8.2kΩ and change the 6.8kΩ 'lower' resistor to 1.0kΩ.

In both cases, trimpot VR1 should have a value of 1kΩ as shown, and should be of the 25-turn cermet type.

Construction

Building the *Precision 10V Reference* is easy. All parts, except the binding post output terminals, switch S1 and the two 9V alkaline batteries are mounted on a single PCB, coded 04104141 and measuring 63 × 53mm. This board fits inside a diecast aluminium box measuring 111 × 60 × 30mm, which not only protects the assembly but also provides shielding.



The PCB is secured to the case lid on two M3 × 15mm spacers at one end before soldering the switch and binding post terminals.

Fig.3 shows the parts layout on the PCB. Note that although trimpot VR1 and its series resistors are shown here, these parts are optional and should only be installed if you can calibrate the device (as mentioned earlier).

Begin the assembly by installing the wire link, then fit the five fixed resistors on the lefthand side of the PCB, plus the two series resistors for trimpot VR1 if it's being used. That done, fit the three multilayer ceramic

Table 2: Capacitor Codes

Value	µF Value	IEC Code	EIA Code
1µF	1µF	1u0	105
100nF	0.1µF	100n	104

Table 1: Resistor Colour Codes

❏	No.	Value	4-Band Code (1%)	5-Band Code (1%)
❏	1	22kΩ	red red orange brown	red red black red brown
❏	1	12kΩ	brown red orange brown	brown red black red brown
❏	2	10kΩ	brown black orange brown	brown black black red brown
❏	1	6.8kΩ	blue grey red brown	blue grey black brown brown
❏	1	2.2kΩ	red red brown	red red black brown brown
❏	1	100Ω	brown black brown brown	brown black black black brown

Parts List

1 diecast aluminium case, 111 × 60 × 30mm
1 PCB, available from the *EPE PCB Service*, code 04104141, 63 × 53mm
1 front-panel label
1 SPST panel-mount momentary pushbutton switch (S1)
1 14-pin DIL IC socket (optional)
1 red binding post terminal
1 black binding post terminal
2 M3 × 15mm tapped spacers
5 M3 × 6mm machine screws
1 M3 hex nut
1 M3 shakeproof washer
2 9V battery clip leads
2 9V alkaline batteries
1 1kΩ cermet trimpot, 25-turn vertical (VR1)
1 100mm length double-sided tape

Semiconductors
1 AD587KNZ or AD587JNZ 10V voltage reference (IC1)
1 4541B programmable CMOS timer (IC2) *
1 BUZ71 or IRF1405 MOSFET (Q1)
2 1N4004 diodes (D1, D2)
1 3mm high-intensity blue LED (LED1)

Capacitors
1 1µF multilayer ceramic
2 100nF multilayer ceramic

Resistors (0.25W, 1%)
1 22kΩ
1 12kΩ
2 10kΩ
1 6.8kΩ (or 1kΩ for 10.240V output)
1 2.2kΩ (or 8.2kΩ for 10.240V output)
1 100Ω

*** see Errata**

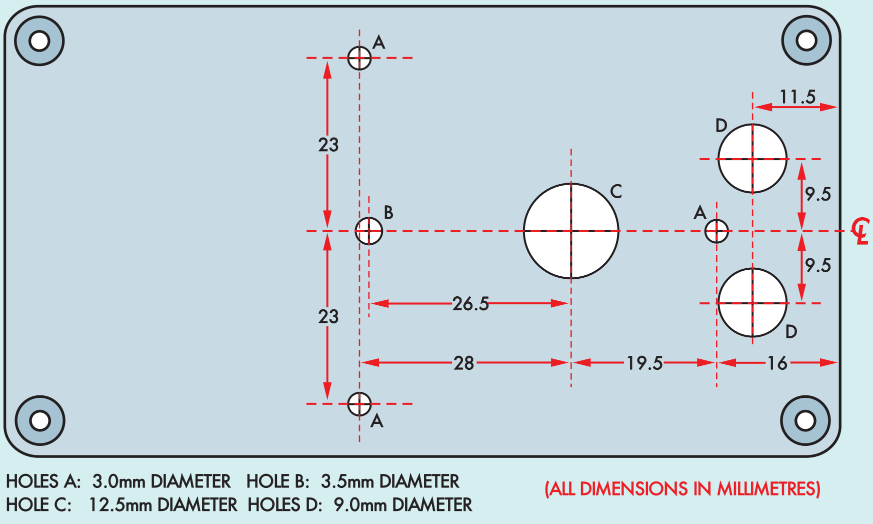


Fig.4: this diagram shows the drilling template for the front panel. It can either be copied or downloaded from the *EPE* website.

capacitors, making sure that the 1µF capacitor goes in at top right.

Now for the two ICs. IC1 must be soldered directly into the board, to ensure reliability (and avoid possible contact resistance). IC2, on the other hand, can either be soldered directly to the PCB or can be installed via a 14-pin DIL socket. Make sure that both ICs are correctly oriented.

Trimpot VR1 is next on the list, followed by MOSFET Q1. Note that Q1's leads must be bent down through 90° about 5mm from its body before mounting it in place. Push it all the way down so that its metal tab sits flush against the PCB and secure it using an M3 × 6mm machine screw, nut and shakeproof washer.

Do the screw up firmly, then solder the MOSFETs leads to their respective pads (note: don't solder the leads first, otherwise the PCB tracks will crack as the mounting screw is tightened down).

LED1 can now be installed, making sure its longer anode (A) lead is oriented as shown. It should be mounted about 7mm proud of the PCB (use a cardboard spacer). Solder just one lead and don't trim the leads at this stage, as you may have to adjust its height later, after the PCB assembly has been mounted on the rear of the lid.

Next, pass the four battery snap leads through the strain-relief holes and solder them to the PCB. That done, cover these connections with silicone to prevent the leads from breaking.

Be sure to connect the red wire from each battery snap to the pad marked '+'.

Your PCB assembly will now be finished and can be placed aside while you prepare the case – or strictly, the case lid, since there are no holes to be drilled in the case itself.

Drilling the case lid

Fig.4 shows the drilling template for the case lid. You have to drill/ream seven holes in all – for the output terminals, switch S1, power LED and PCB mounting, plus a screwdriver access hole for trimpot VR1 (if necessary).

Fig.4 shows the location and size of each of these holes. You can either follow this diagram to mark out the lid for drilling – or you can copy it, cut it to size and attach it directly to the lid (using double-sided tape) for use as a drilling template. The drilling template is also available for download from the *EPE* website.

Use a small pilot drill to start the holes, then remove the template and carefully drill and ream them to size. Deburr each hole with an oversize drill or in the case of the three larger holes, a small rat-tail file.

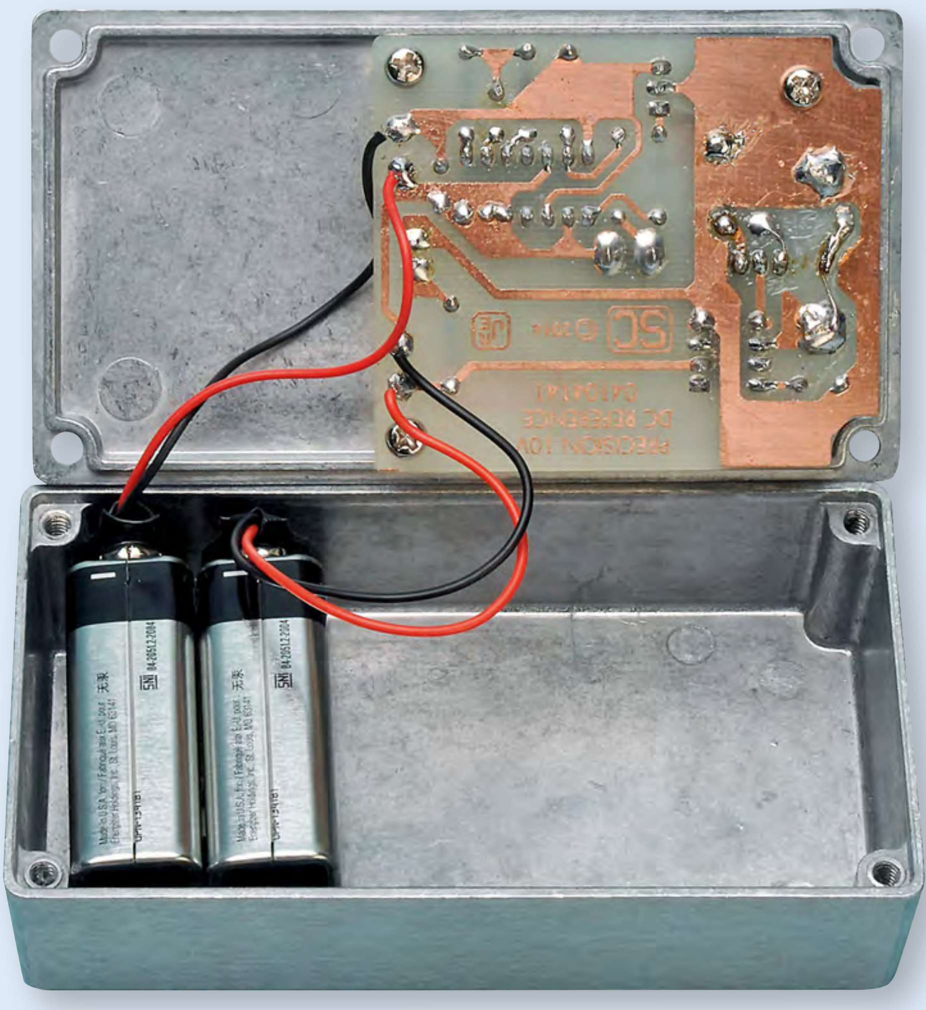
Now for the front panel artwork. This artwork can be obtained either by photocopying Fig.5 onto an adhesive-backed label, or it can be downloaded as a PDF file from the *EPE* website and printed out. It can then be covered with a self-adhesive transparent film to protect it from finger marks.

Alternatively, it can be photocopied onto plain paper, hot-laminated into a clear protective sleeve and then attached to the lid using double-side tape or silicone adhesive. The various holes can then be cut out using a sharp hobby knife.

Pushbutton switch S1 can now be mounted on the lid, taking care to orient it so that its two connection lugs are aligned along the long axis. This is necessary so they will later fit through their holes in the centre of the PCB. That done, attach the two output terminals (binding posts) to the lid, making sure that the red terminal goes to the '+' position and the black terminal to the '–' position.

Specifications

- Output voltage: 10.000V DC (10.240V optional – see text)
- Basic accuracy: ±0.05% (±5mV) without adjustment, ±0.002% after trim adjustment
- Long term drift: <15ppm per 1000 hours, mostly in first year of operation
- Temperature stability: <7mV change between 0°C and +70°C
- Maximum output current: 10mA
- Noise on output: <4μV peak-to-peak (0.1Hz – 10Hz); <180μV peak-to-peak (DC – 1MHz)
- Load regulation: less than ±100μV/mA for loads up to 10mA
- Power supply: 2 x 9V alkaline batteries; quiescent current drain (when operating) <6.5mA
- Auto-off time: 90 seconds; standby current 10nA



This is the view inside the completed unit. The two 9V batteries are held together and to the bottom of the case using double-sided adhesive tape.

Tighten the mounting nuts of the terminals as firmly as possible, so that they're held securely in place.

Final assembly

As shown in the photos, the PCB assembly mounts on the back of the lid

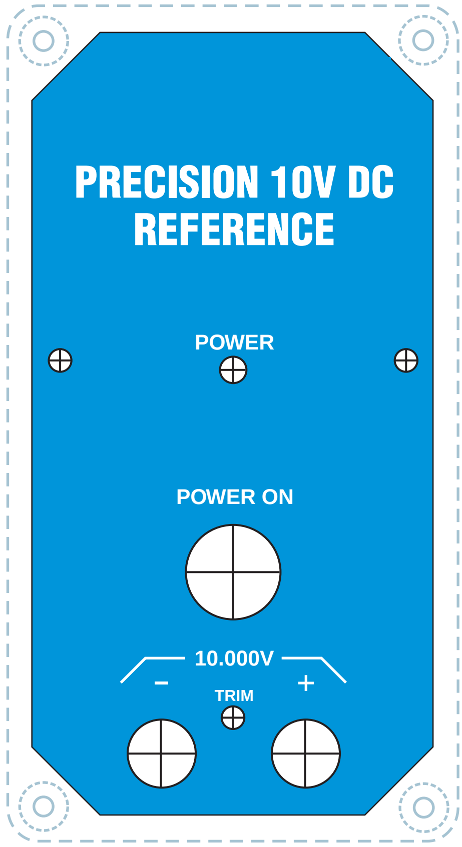


Fig.5: this full-size front panel artwork can be laminated and attached to the case lid using silicone adhesive or double-sided tape.

and is supported by two M3 × 15mm tapped spacers at one end and by the two output terminal connections at the other.

The first step is to fit the two M3 × 15mm spacers to the 'battery end' of the PCB. That done, the PCB can be fitted in place, making sure that:

1) Both switch lugs pass through their matching holes
2) LED1 passes up through its corresponding hole in the lid
3) The binding post spigots pass down through their matching holes in the PCB. The PCB can then be fastened in position using two more M3 × 6mm machine screws which pass through the lid and into the spacers.

Once it's in place, the switch lugs and binding post spigots can be soldered to their respective PCB pads. If necessary, the solder connection on the LED lead can then be melted and the LED adjusted so that it just protrudes through its front-panel mounting hole. The remaining LED lead can then be soldered and the first lead connection remade with some fresh solder.

Device availability

Analog Devices make 18 different versions of the AD587, many of them in small outline (SOIC) SMD plastic or CERDIP packages. By contrast, the AD587KNZ and AD587JNZ both come in 8-pin PDIP packages and are quite reasonably priced; see below for spring 2015 prices.

Both are currently available from suppliers such as element14 and RS Components. For example, element14 (http://uk.farnell.com) has the AD587KNZ (order code 2143134) available for £6.60 plus VAT, while the lower-spec AD587JNZ (order code 9605169) costs £7.15 plus VAT.

Similarly, RS Components (http://uk.rs-online.com) sells the AD587KNZ (order code 523-7415) for £4.58 plus VAT, while the AD587JNZ (order code 412-579) is slightly less at £4.02 plus VAT.

You shouldn't have any trouble getting the 4541B programmable timer, either. For example, RS has it (order code 709-1999) for £0.27 (in packs of 10).

Finally, the battery snap leads can be fitted to a pair of new 9V alkaline batteries, after which the batteries can be held together using a strip of double-sided adhesive tape between them.

Two more strips of double-sided tape are then used to secure the batteries to the bottom of the case, after which the lid/PCB assembly can be fitted and the lid fastened down using the four countersunk M4 screws supplied.

That's it – your *Precision 10V DC Reference Mk.2* is complete. So, now it's time for the smoke test!

Using it

There are no adjustments to be made to the unit, unless (as stated) you have a high-precision, recently-calibrated DMM to calibrate it against. If you're not calibrating the unit, you will be relying on the ±5mV or better precision provided by the AD587KNZ chip itself. In that case, check that VR1 and/or its two associated resistors have been left out of circuit, otherwise the accuracy of the unit will be compromised.

Using the *Precision 10V Reference* is simple – just press S1 to turn the the unit on for about 90s. As soon as you press S1, LED1 should light to show that the unit is operating and providing 10.000V ±5mV at its output terminals, ready for calibrating your DMM or whatever.

If you haven't finished making measurements when LED1 turns off (ie, when the unit unit powers down), it's simply a matter of pressing S1 again to power it up for another 90s.

Incidentally, you'll find that when you first connect the battery snap leads to the batteries, LED1 will turn on to show that the unit is operating. This is normal and is simply due to the way that the 4541B timer chip works.

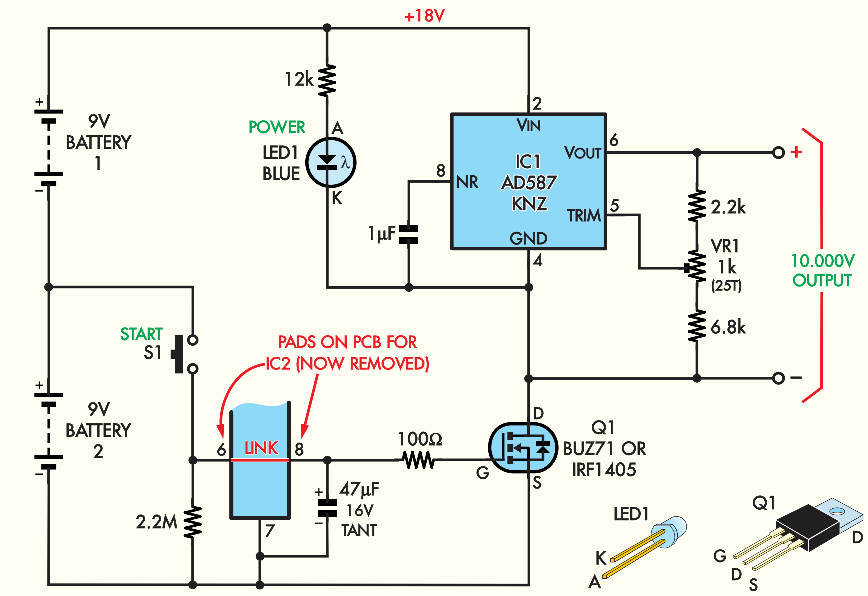
Finally, if you wish to calibrate the unit, make sure VR1 and its associated resistors have been installed. It's then just a matter of monitoring the output on a 6.5-digit (or better) bench DMM and adjusting VR1 to get a reading as close as possible to 10.00000V (or 10.24000V if you prefer).

Errata: solving the current-drain problem

Not long after the original *10V Precision Voltage Reference* was published in *Silicon Chip* we received emails from several readers, complaining about higher than expected current drain causing one of the batteries to go flat prematurely. This was because the timing oscillator in the 4541B digital timing chip continues to run even when the chip has not been triggered, drawing a current of around 1mA from the lower 9V battery even when the device is nominally 'off'.

This drawback is not apparent from the 4514B data sheet (as far as we could see) and we tried a number of measures to disable the timing oscillator when the chip is in the triggered state; none were successful.

Ultimately, we have solved the problem by removing the 4514B and its associated timing components and using an RC network to provide the timing period for MOSFET Q1. Pins 6 and 8 of the vacant IC2 position are bridged and a 47µF tantalum capacitor is connected between pins 7 and



8, and a 2.2MΩ resistor is connected between pin 6 and the 0V line.

The resistor and capacitor then form an RC time-delay circuit which keeps MOSFET Q1 biased 'on' for about 110 seconds after pushbutton switch S1 (Start) is released. The timing may not be as precise as provided by the 4514B originally used, but it does have the merit of zero current consumption when MOSFET Q1 turns off.

Jim Rowe
Project designer

Be the Conductor!

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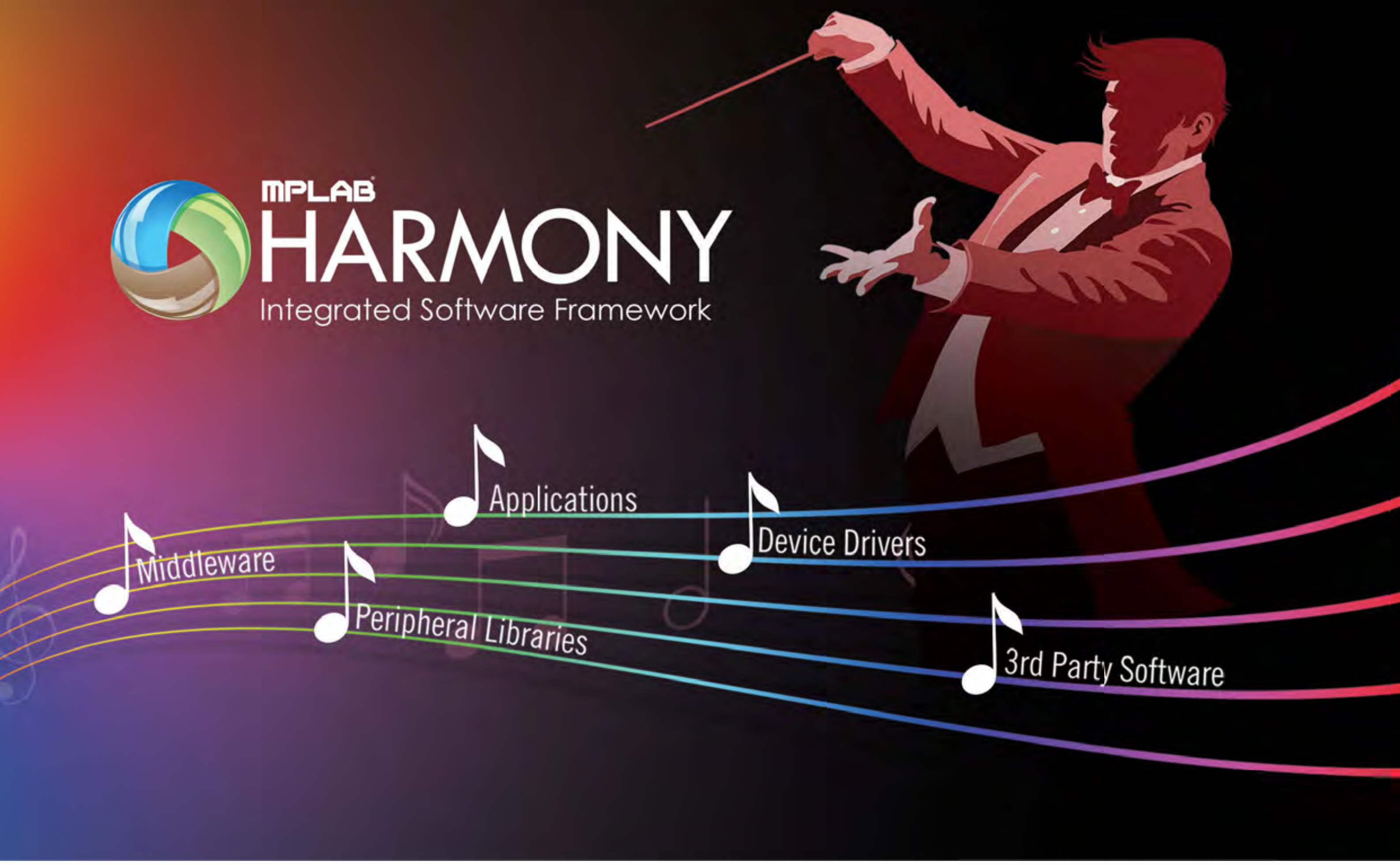

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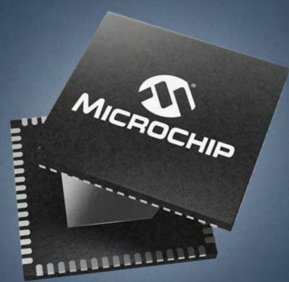

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TechnoTalk

Mark Nelson

For a long time 'the Internet of Things' sounded like pretentious nonsense, of little relevance to normal people – you and me. In fact, it's creeping up on us rapidly, as three new applications will prove. They're practical, electronic and affect our everyday life, hence highly relevant to this magazine. Mark Nelson explains.

Start with a rant!

'New in box', 'shrink wrapped' and 'sealed wrapping' are three expressions that seem to have taken on new meanings on well-known auction sites. You'd be forgiven for assuming you were bidding on unopened, brand-new goods – but of course that would be sheer *naiveté*. This happened to a friend of mine a fortnight ago, who ranted: 'He sold me a CD that was described as "sealed in cellophane" and then got chippy when I posted feedback for him saying that "it was sealed in cellophane, but not the original cellophane". In other words, he's bagging used CDs and using this wording to suggest he's supplying something he isn't.'

When challenged, the vendor replied in weasel words: 'Your CD was listed as "like new", not "new", so the cellophane was never going to be the original. If it were, I would have listed it as "factory sealed". I routinely rewrap all my second hand CDs after putting them in new jewel cases. This is not an attempt to deceive, as they are not listed as new, but it protects the new jewel cases from getting scratched and keeps moisture out.'

Yeah, right. In any case, has anyone seen anything wrapped in proper cellophane during the last 50 years? People use plastic – not cellulose – film these days.

OpenSense

Back in the real world, commercial sellers rely on plastic film wrapping to guarantee the integrity of the products they are selling. The slightest suggestion that foodstuffs, drugs or patent remedies on shop shelves might have been tampered with puts the media into a state of panic, so an automatic method to detect meddling would be the answer to every retailer's prayer. Well, that prayer has just been answered with OpenSense, a new near-field communication (NFC) sensor tag technology that promises to enhance consumer confidence and improve product security. Suggested applications include wine and spirits, pharmaceuticals, cosmetics, health and beauty care, and even safety-critical automotive parts.

Developed in California for the Scandinavian firm Thinfilm Electronics, OpenSense enables the dynamic detection of a product's 'sealed' and 'open' states, with smartphone-enabled NFC readability of the tag before and after the factory seal has been broken. With a simple wave of an NFC-enabled smartphone or mobile device, consumers, branded product manufacturers, and trusted partners can instantly receive relevant messaging. Secure messaging also provides robust supply-chain analytics to enable smart business decisions (and discourage theft or even make it virtually impossible).

Scary? Daft? Over-sophisticated? – not according to Thinfilm. Says Davor Sutija, Thinfilm's chief executive officer, 'The interest we've received to date has been tremendous. Ultra-high production levels and relatively lower unit costs can help facilitate wider, item-level deployments of OpenSense, greatly supporting Thinfilm in its mission to make everyday objects smart and extend the boundaries of the Internet of Things (IoT).'

Boxy but safe

We all know that cars can't talk (except in cartoon films). Even the nodding dog in the rear window has only two words, 'Oh' and 'Yes'. But Volvo cars ('They're boxy but safe' as the spoof slogan said) are now chatting like mad in Sweden and Norway. A thousand Volvo cars are involved in a cloud-based project to enable cars to share information about conditions that reduce road grip (such as icy patches). This is a joint collaboration between Volvo Cars, the Swedish Transport Administration and the Norwegian Public Roads Authority, with a goal of making the technology available to customers within a few years' time.

'The more information that can be shared on the road, the fewer surprises there are. And when you're driving, surprises are what you most want to avoid,' says Volvo's Erik Israelsson. 'In light of that, we've developed a slippery-road alert, which notifies drivers about icy patches and contributes to making winter road maintenance more efficient. We're also adding a hazard-light alert, which will tell drivers if another vehicle in the area has its hazard lights on. With these first two features, we have a great platform for developing additional safety features. This is just the beginning.'

The test area is being broadened to include two major Scandinavian cities, Gothenburg and Oslo, so as to provide a more complete picture of how the system will work in real winter traffic conditions. 'This will bring us closer to our safety vision that by 2020 no one should be killed or seriously injured in a new Volvo car. And it's another way in which the "Designed around you" philosophy improves the driving experience,' he adds.

Parking pain relieved

According to car parks operator Apcoa, motorists drive an average of 2.8 miles before they finally find somewhere to park. But a new sensor-controlled parking management system from Siemens is helping to optimise the use of on-street parking spots and radically reduce the congestion caused by motorists searching for a space. A real-world pilot project is scheduled this summer in Munich, Germany.

Radar sensors mounted on lamp posts monitor parking place occupancy, using Intel's Internet of Things platform to connect securely and flexibly with the control centre. The system also uses RFID transponders to identify resident parking permits, special permits for disabled persons, vehicles that are part of a car sharing system, electric vehicles and similar priority users. This technology could also be used for automated payment solutions in future. What is not explained is how car drivers exploit this information. The publicity photo shows a driver scrutinising a map displayed on a tablet computer balanced above the dashboard – and not watching the road ahead. Perhaps it's not the smartest idea after all. The concept is ingenious, but the implementation looks lethal and relies on users having a tablet with good mobile coverage, something that plenty of drivers are unlikely to possess.

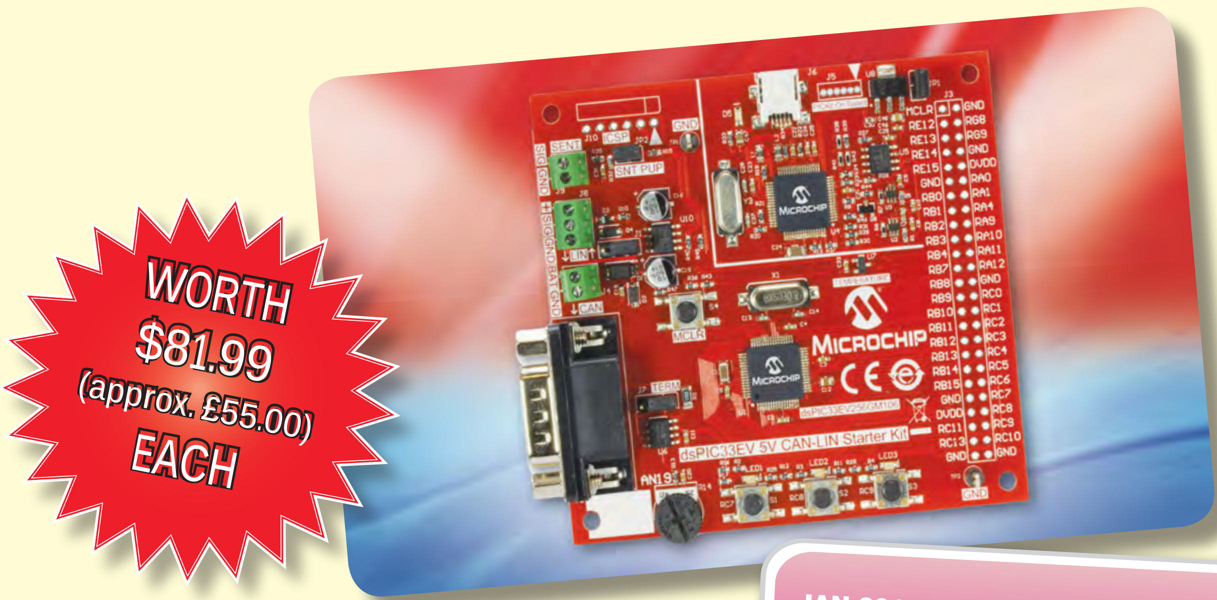
Win a Microchip dsPIC33EV 5V CAN-LIN Starter Kit

EVERYDAY PRACTICAL ELECTRONICS is offering its readers the chance to win a Microchip dsPIC33EV 5V CAN-LIN Starter Kit! The Starter Kit (DM330018) features the dsPIC33EV256GM106 Digital Signal Controller (DSC) for automotive and motor control applications.

The Starter Kit contains serial data ports for CAN, LIN and SENT, a self-contained USB programming/debug interface, and an expansion footprint for flexibility in application hardware development. This board allows users to explore three popular automotive and industrial serial data formats (CAN, LIN and SENT). The PICkit On-Board (PKOB) USB programmer and debugger allows simple programming without the need for an additional hardware interface.

With 5V operation up to 150°C, Microchip's dsPIC33 'EV' digital signal controllers (DSCs) provide harsh applications with robust performance as well as integrating peripherals for safety-critical functions, motor control, CAN, SENT and touch sensing.

The dsPIC33 'EV' family is the first to offer error correcting code (ECC) Flash for increased reliability and safety. For safety-critical applications, cyclic redundancy check (CRC), deadman timer (DMT), windowed watchdog timer (WWDT) peripherals as well as a backup system oscillator and certified Class B software.



JAN 2015 ISSUE WINNER

Mr André Smets who works at Dimension Data, Belgium.

He won a PICDEM Lab Development Kit valued at $134.99.

HOW TO ENTER

For the chance to win a Microchip dsPIC33EV 5V CAN-LIN Starter Kit, please visit: **http://www.microchip-comps.com/epe-dspic33ev-can-lin** and enter your details onto the online entry form.

CLOSING DATE

The closing date for this offer is 31 May 2015

Deluxe Fan Speed Controller

By
John Clarke

- Full Range: slow to maximum speed control
- Suits ceiling or plug-in fans
- No mysterious fan noises in the night
- No AM radio interference

Got a ceiling fan or pedestal fan? With limited speed settings they are often too fast or two slow. This non-switched controller gives you continuous speed control and as a bonus, it produces no radio interference or motor noise. It can also be used as a dimmer for desk and reading lamps up to 60W.

Ceiling fans usually offer just three switched speed settings: too fast, fast and not slow enough. We decided to produce a controller which gives a wide range of speeds, from the maximum down to quite low, to give just the faintest of breezes.

But we decided to *not* take the obvious approach of using a phase-controlled triac to produce the speed control because they can cause considerable interference to AM radio reception, particularly in those areas where signals are weak.

Instead, our controller is based on a high-voltage MOSFET, which is effectively a variable resistor connected in series with the fan motor. For high fan speeds, the MOSFET resistance is low and for lower speeds, the MOSFET resistance is higher.

And while we have only mentioned ceiling fans up to this point, it can also be used with pedestal or table fans. You simply plug the fan into a switched mains socket on the controller's case lid, while the controller plugs into the mains via an IEC mains lead.

If you have a ceiling fan, it may need to be rewired permanently (by a licensed electrician).

Since the speed control element is essentially a variable resistor, it will not be very efficient in electrical terms and that means it will dissipate some heat. But considering that most fans will draw only up to about 60W at full speed and less as speed is reduced, the dissipation can be managed by using a diecast box and finned heatsink.

Bear in mind that we don't need to dissipate anywhere near 60W – at full speed the dissipation in the controller

Features

- Full control of motor speed from stopped to maximum
- For 230VAC shaded pole and capaitor-run motors
- Over-current limiting
- Over-temperature cut out
- Quiet operation
- Fused circuit
- Rugged case

is quite small. It's at lower speeds that dissipation in the controller increases.

But because the motor is running slower, overall power is less than at full speed. If it does get too hot, there is an over-temperature thermostat to switch the controller off.

We cannot connect the high-voltage MOSFET directly in series with the 230VAC mains supply to the fan because MOSFETs can only work from DC, or at worst, from fluctuating DC. Any reverse current would be shunted by the MOSFET's intrinsic internal diode – so that wouldn't work.

The solution is quite simple though; we use a bridge rectifier. That way, the MOSFET is only subjected to rectified AC (or fluctuating DC) yet it can comfortably control the AC load of the fan.

Fig.1 shows the general arrangement. The MOSFET and current-sensing resistor connect between the positive and negative terminals of the bridge rectifier. When the live voltage is more positive than the neutral, current (i1) flows through the motor, diode D1 and through the MOSFET from plus to minus of the bridge, then through D3 and to neutral.

When the live is more negative than the neutral, current (i2) flows from the neutral through D4 and the MOSFET from the positive to the negative of the bridge and then through D2 and the motor to live. The current flow through the resistive element is therefore always from the positive to the negative terminals of the bridge rectifier.

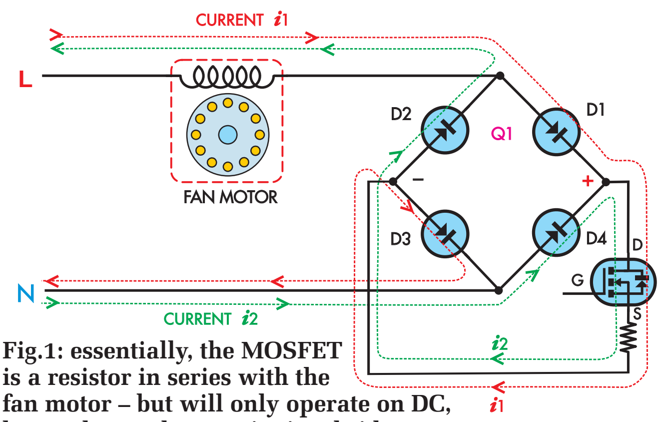


Fig.1: essentially, the MOSFET is a resistor in series with the fan motor – but will only operate on DC, hence the need to run it via a bridge rectifier. Current i1 and i2 are the two halves of the AC waveform, so the motor is still fed with AC.

Circuit description

The circuit for the *Fan Speed Controller* is shown in Fig.2. It comprises just one IC, several diodes, the high voltage MOSFET, Q1, plus some resistors and capacitors.

The circuit and wiring diagram are for free-standing fans (ie, those connected via a 3-pin plug). For ceiling fans, some components are not required - we'll look at these later.

Power for the circuit is derived directly from the 230VAC mains. **The entire circuit floats at mains potential, so it is unsafe to touch whenever the circuit is connected to the mains**. Additionally, the circuit ground is also floating at mains potential and is not connected to mains earth. **The metal box housing the controller is connected to the mains earth**.

Mains power is supplied to the controller circuit via an IEC socket and fuse, F1, which is an integral part of the IEC input connector. Fuse operation protects the circuit against excessive current flow should a fault occur, such as a short across the motor.

BR1 is a 6A bridge rectifier with a 400V rating. As mentioned, the bridge provides the MOSFET with the positive full-wave rectified mains voltage, while the fan motor receives AC.

A separate supply is provided for the low voltage circuitry. We use another bridge rectifier (BR2) and derive a low voltage supply via 220nF capacitors from the 230V mains. The capacitors are in preference to high wattage resistors since they do not dissipate significant power, therefore reducing heat dissipation inside the controller case.

The circuit shows the arrangement with the separate rectifier (BR2) fed via two 220nF capacitors and series 470Ω resistors.

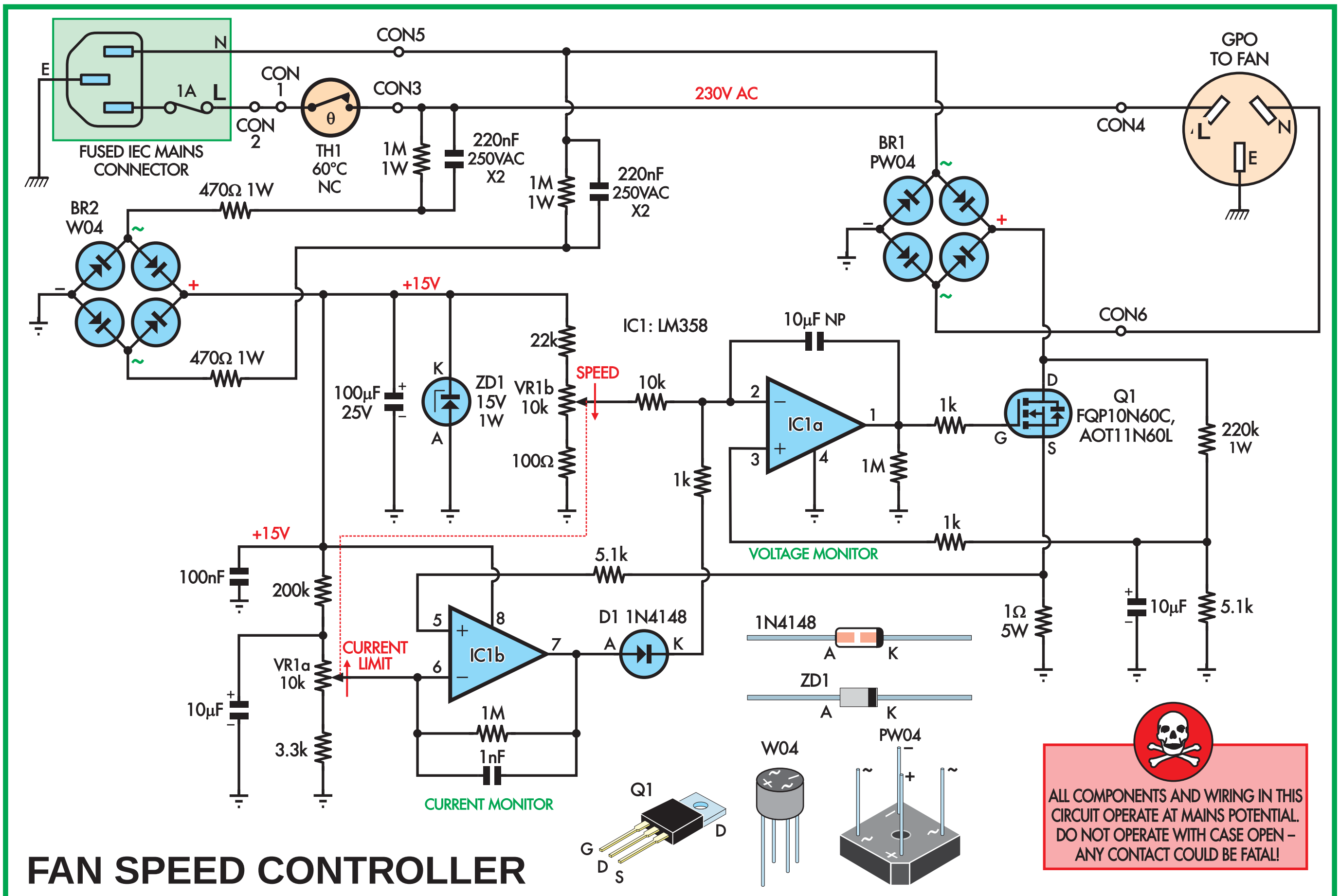


Fig.2: the circuit for our new *Fan Speed Controller* shows it has two bridge rectifiers, one of which provides low voltage DC direct from the mains. This is used to power the rest of the circuit. The second bridge (BR1) allows a power MOSFET to control the current to the AC motor over both halves of the 230V mains cycle. The MOSFET acts like a variable resistor, supplying more or less power to the fan motor depending on the setting of VR1a and b.

The 220nF capacitors provide an impedance that limits current flow to the 15V Zener diode ZD1. At 50Hz, the impedance of each 220nF capacitor is 14.5kΩ. This impedance plus the 470Ω limits current to the 15V Zener diode, ZD1 to about 10mA. A 100µF capacitor across the resulting 15V supply smooths it to a constant DC voltage.

The 470Ω resistors in series with the 220nF capacitors are there to limit surge current when power is first applied to the circuit. The surge current could be high should power be switched on at the peak voltage of the mains waveform. 1MΩ resistors across the capacitors are to discharge them when the power is switched off.

The 15V supply powers the LM358 dual op amp, IC1. One of these operational amplifiers, IC1a, is used to drive the gate of MOSFET Q1. This op amp is connected in a feedback control loop that monitors both a divided version of the voltage between Q1's drain and source and the voltage provided by speed potentiometer VR1b. IC1a adjusts its output voltage at the MOSFET gate so that the divided drain-source voltage across the MOSFET matches that set by the speed potentiometer.

In more detail, a 220kΩ 1W resistor and a 5.1kΩ resistor form a voltage divider across Q1 (ignoring the series 1Ω resistor). This effectively reduces the voltage across Q1 to about 1/44 its original value, calculated as (5.1k + 220k) ÷ 5.1k. The resulting voltage is filtered with a 10µF capacitor providing a DC voltage from the full-wave rectified waveform.

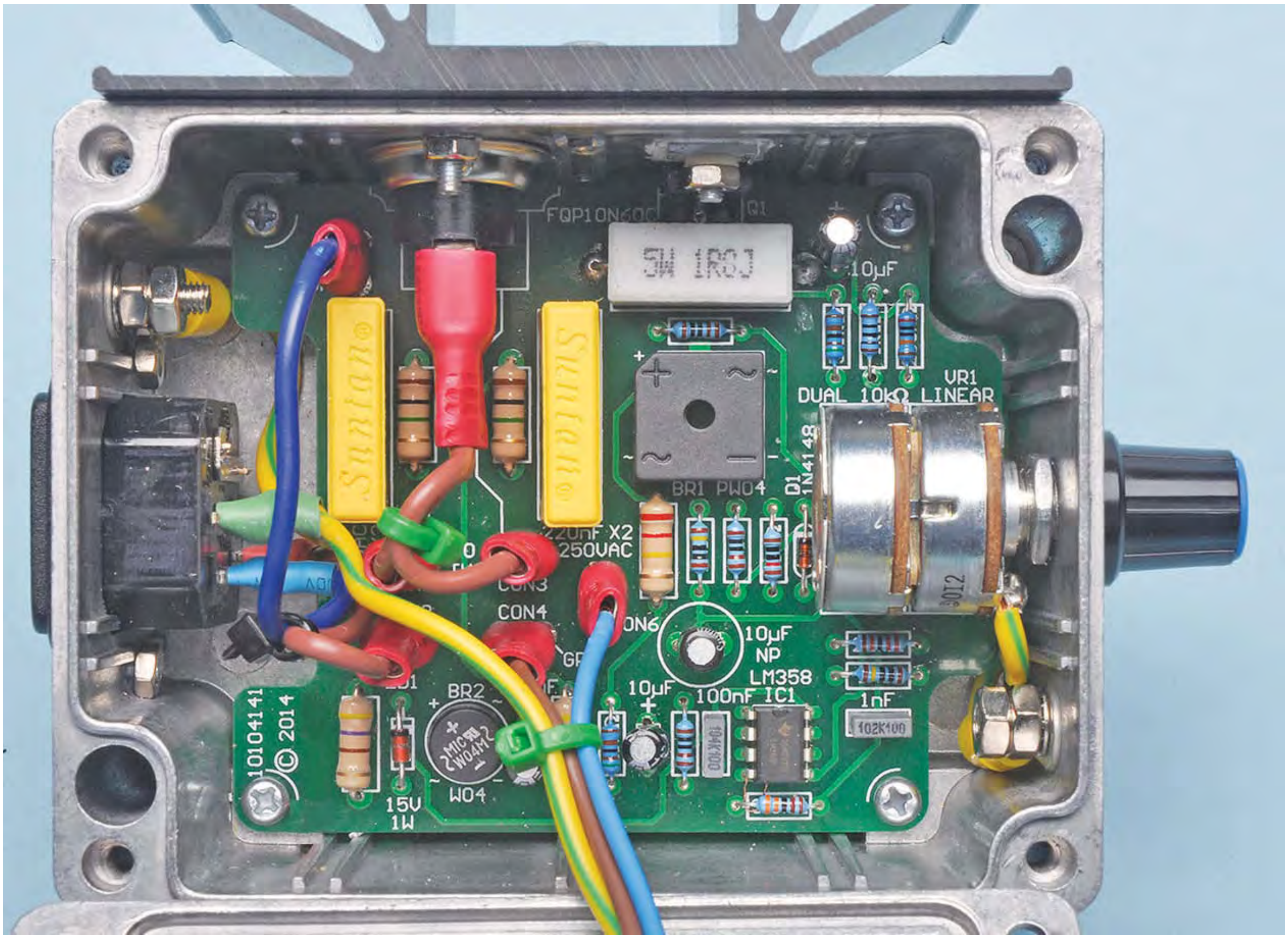
The resistive divider is there to produce a suitable low voltage for monitoring by IC1a. The maximum voltage needs to be several volts below the positive supply for IC1 at 15V. That's because the op amp is designed to operate with inputs that can go down to the negative supply, but not as high as the positive supply.

Maximum voltage from the divider occurs when Q1 is at a high resistance. Then the full 230VAC of the mains supply is across the MOSFET. The peak of the 230V RMS waveform is 325V and after reduction by a factor of 44, brings the voltage down to 7.39V peak. This becomes 4.7V DC after filtering with the 10µF capacitor. Note that this average voltage of the full-wave rectified waveform is 0.63 of the waveform peak.

As the resistance of Q1 is decreased, there is more voltage across the fan motor and less across the MOSFET. The voltage from the divider is therefore also lower.

VR1b is the speed control adjustment. VR1b is connected in series between a 22kΩ resistor from the +15V supply and a 100Ω resistor connecting to the 0V supply. With this resistor string, the voltage range for the wiper of VR1b is between 5V and 0.05V.

Operation is as follows: if VR1b is set to produce, for example, 2V DC at its wiper – IC1a adjusts its drive to the



Looking inside the open 'IP65' case shows how easy the PCB mounts on the tapped supports inside. Note that we do not have the IEC power lead plugged in – neither should you whenever the case is open!

gate of Q1 so that the voltage monitored at the divide-by-44 resistors is also 2V DC. With 2V on the divider it means that there is 88V (average) across Q1.

The 88V average is equivalent to 97.5V RMS. If the mains voltage is at 230VAC RMS then the voltage across the fan is: 230V – 97.5V = 132.5V RMS.

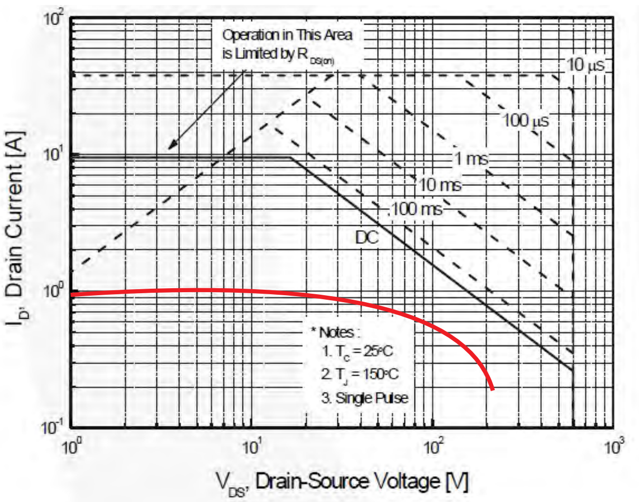


Fig. 3: SOA graph for the FQP10N60C MOSFET used in this project. The text explains how to interpret this.

Note that for VR1b, the lower voltage is deliberately made to be slightly above 0V using the 100Ω resistor. This is to prevent IC1a from oscillation at the lowest voltage position for VR1b.

The voltage feedback control ensures that voltage across the MOSFET is strictly maintained to prevent changes in the motor speed. That's provided the mains voltage remains reasonably constant (which it usually does). Without the feedback control and just applying a fixed voltage to the gate of Q1, the fan would slow quite markedly as the MOSFET heats up. That's because the MOSFET drain to source resistance increases with temperature.

Current limit

Current limiting for this circuit is necessary due to the fact that while the MOSFET can happily conduct around 10A, this is only when there is a relatively low voltage between its drain and source. With a high voltage between drain and source, the current needs to be reduced to prevent internal damage to the MOSFET.

Incidentally, no domestic fan (plug-in or ceiling) will demand anything like 10A. They're much more likely to draw a tiny fraction of this – most fans are rated at 10-50W, which equates to just 40-220mA!

Fig.4 shows the 'safe operating area' (SOA) of the FQP10N60C MOSFET. The lower DC, SOA line shows

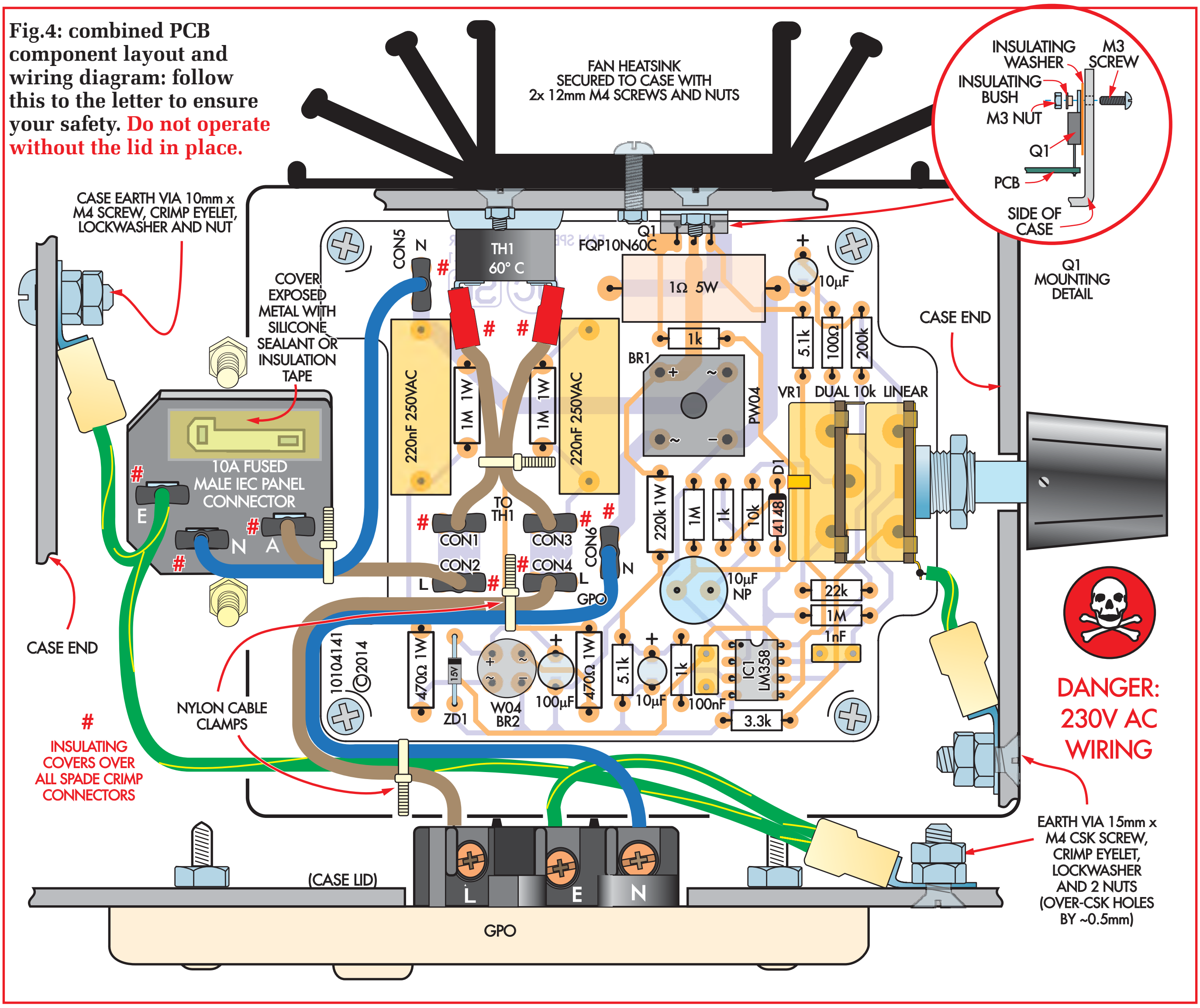


Fig.4: combined PCB component layout and wiring diagram: follow this to the letter to ensure your safety. Do not operate without the lid in place.

that the device can easily supply up to 10A to the fan motor but as the drain-to-source voltage increases above around 20V, the MOSFET current rating falls, to 800mA at 200V.

The red line indicates the current limit our circuit applies to safeguard the MOSFET from exceeding the SOA. We restrict the maximum current to around 1A up to around 20V between drain and source. At this drain-to-source voltage, the fan will run at a fast speed. At lower fan speed settings, the voltage between the drain and source will be higher and we limit the current to prevent this exceeding the SOA curve. For the slowest speeds the current is limited to around 230mA.

Note that this SOA curve is for the non-insulated MOSFET package. For fully insulated MOSFET packages (eg, FQPF10N60C) both the SOA curve and thermal resistance from junction to case is worse. The thermal resistance is some three times higher. It means the insulated package, while more convenient for mounting, is unsuited for this application. The MOSFET would overheat internally, regardless of the amount of heatsinking.

Additionally for the insulated package, for SOA, the 10A current rating is only for up to 5.5V drain to source. For these reasons we use the non-insulated MOSFET package.

IC1b provides the current-limit function. It monitors the voltage across the 1Ω 5W resistor that is in series with Q1. The 1Ω resistor converts the fan current to a voltage. A 1A current, for example, will result in 1V across this resistor. IC1b is connected as an amplifier that has level shifting set by VR1a. As the voltage across the 1Ω resistor exceeds the voltage set at the wiper of VR1a, the IC1b output goes high and drives the input pin 2 of IC1b high via diode D1 and the 1kΩ series resistor.

This over-rides the motor speed setting, slowing fan speed to reduce current. If the current monitor voltage from the 1Ω resistor is less than the voltage set at the wiper of VR1a, IC1b output is low and thus has no effect on IC1a as diode D1 is reverse biased. VR1a is connected across the 15V supply in a similar way to VR1b, but the upper and lower resistors are different values. The 200kΩ and 3.3kΩ resistors set the VR1a current-limit range to between 940mV and 235mV. Both VR1a and VR1b are physically connected to one potentiometer shaft, so adjusting fan speed will also automatically adjust the current limit.

Construction

With the exception of the mains input and output sockets and thermal cutout, all components mount on a single PCB

which is available from the *EPE PCB Service*, coded 10104141, measuring 93 × 79mm. It is designed to be housed in an IP65 diecast box measuring 115 × 90 × 55mm.

The PCB is shaped to match the internal contours of the IP65 case and has a cutout to fit the IEC input connector.

However, this case is relatively expensive – if you wish, the *Fan Speed Controller* can be built into a (slightly larger) economy diecast case instead. The PCB will then need to be mounted onto separate standoffs with four holes drilled in the base for these.

Begin construction by checking the PCB. We do not expect any problems with PCBs as supplied by the *EPE PCB Service*, or with those supplied in kits. These are of high quality and are solder masked, screen printed and shaped with the required cut outs.

It is still worthwhile to check if there are problems with the PCB and look for any shorts or breaks between tracks. If there are any problems, repair these as necessary. Similarly, if the cut outs in the sides of the PCB have not been shaped, they should be cut and filed to size before any components are assembled.

Check that the PCB fits into the case before starting assembly. With the IP65 case specified, the PCB conveniently mounts on the integral tapped lands provided.

Follow the overlay diagram shown in Fig.4. Begin by soldering in the resistors, using the accompanying table for the colour codes. Diode D1 can be inserted next, taking care to orient it correctly.

IC1 can be directly mounted or you can use an IC socket. Either way, be sure to install the socket and/or the IC the correct way around with the notch facing the direction shown on the overlay.

Capacitors can be installed next. The accompanying capacitor table shows the various codes that are used to indicate the capacitance values of the MKT polyester and X2 class capacitors. The electrolytic capacitors have their value directly marked and the polarised types must be oriented correctly.

The NP capacitor can be mounted either way.

You can use 10µF ceramic surface-mount capacitors instead of the electrolytic types if you wish, and provision has been made for these on of the PCB. If using these, position and tack-solder each in place, making sure they are

Parts List – Deluxe Fan Speed Controller

1 PCB, available from the *EPE PCB Service*, coded 10104141, 93 × 79mm
1 IP65 diecast box measuring 115 × 90 × 55mm
1 lid and side panel label
1 fan type heatsink 105 × 25.5 × 55mm
1 Architrave GPO outlet (Clipsal CLI16WE or equivalent)* – UK readers need a UK version
1 Male IEC mains connector with integral M205 fuseholder
1 1A M205 fuse
1 7.5A IEC mains lead
1 10A thermostat 60°C normally closed (NC)
1 10k dual ganged 24mm PCB-mount linear pot (VR1)
1 plastic knob to suit potentiometer shaft
6 6.35mm PCB-mount male spade connectors, 5.08mm pin spacing (CON 1-6)
8 6.35mm insulated female spade quick connectors for 1mm wire diameter (red)
3 5.3mm ID insulated quick connect crimp eyelets for 2-5mm wire diameter (yellow)
2 M4 × 15mm countersunk screws (lid and potentiometer side earth)
3 M4 × 15mm screws (GPO and IEC end earth)
2 M4 × 10mm screws (securing heatsink when the case is M4 tapped) (use 2 M4 × 15mm screws and two extra M4 nuts when case is not M4 tapped)
8 M4 nuts
5 4mm star washers
2 M3.5 × 6mm screws (for PCB mounting) [in addition to the two supplied with case]
3 M3 × 10mm countersunk screws (for Q1 and TH1)
2 M3 × 10mm countersunk screws (for IEC connector)
5 M3 nuts
1 TO-220 Mica insulating washer
1 TO-220 insulating bush
4 small stick on rubber feet
1 200mm length of green/yellow 7.5A mains wire
1 200mm length of brown 7.5A mains wire
1 200mm length of blue 7.5A mains wire
1 70mm length of 5mm heatshrink tubing
4 100mm cable ties
Heatsink compound

Semiconductors
1 LM358 DIP dual op amp (IC1)
1 600V 9A or more N-channel MOSFET (FQP10N60C, AOT11N60L, BUK457-600B) (Q1)
1 15V 1W Zener diode (ZD1)
1 400V 6A P04 diode bridge (BR1)
1 400V 1.2A W04 diode bridge (BR2)
1 1N4148 signal diode (D1)

Capacitors
1 100µF 105°C 16V PC electrolytic*
1 10µF 105°C 50V NP PC electrolytic*
2 10µF 105°C 16V PC electrolytic*
2 220nF 250VAC X2 class
1 100nF 63V or 100V MKT Polyester
1 1nF 63V or 100V MKT polyester

Resistors
(0.25W, 1%)

2 1MΩ	1 200kΩ	1 22kΩ	1 10kΩ
2 5.1kΩ	1 3.3kΩ	3 1kΩ	1 100Ω

(1W, 5%)

2 1MΩ	1 220kΩ	2 470Ω
1 1Ω 5W		

*Notes:
An economy diecast box 119 × 94 × 57mm can be used instead of the IP65 case; a larger box may be required depending on the type of mains socket used.

Extra parts required include 4 6.3mm M3 tapped standoffs and 8 M3 × 5mm screws. The two M3.5 × 6mm screws are not required.

All the 10µF (polarised and NP types) and the 100µF electrolytic capacitors can each be replaced by 10µF surface mount ceramic capacitors (10µF 50V 3216 (metric)/1206 (imperial)). These will provide a longer service life than electrolytic capacitors. Provision has been made to mount these where each electrolytic capacitor would normally be positioned. Ceramic capacitors are not polarised so can be oriented either way on the PCB.

aligned correctly before soldering them fully in place.

Diode bridges BR1 and BR2 can be installed next, take care to orient these correctly and in the right locations.

Before installing VR1, its shaft may need to be cut to length to suit the knob. The potentiometer nut is wound fully onto its thread. This nut is adjusted later to make contact with the inside of the case.

Finally, install the PCB spade connectors at CON1-CON6.

Mounting the hardware

A marking-out guide and panel artwork are provided on the *EPE* website. This provides the IEC connector and GPO cut outs for the end and front panels (different cut outs may be required for various mains sockets). Details are given for both the IP65 and economy box.

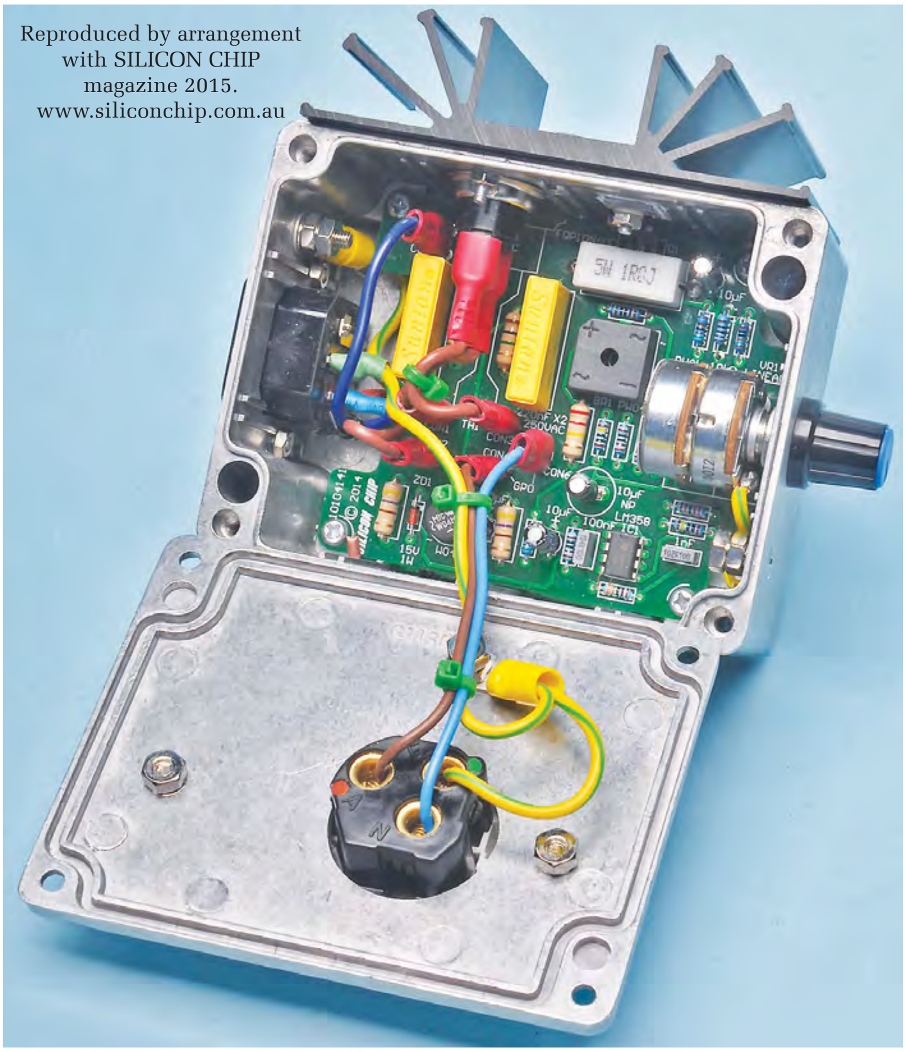
First of all, mark out the hole position for the IEC connector and earth screw in the end wall of the case. There is about a 4mm gap from the base of the case to the bottom of the IEC connector. The hole is made by drilling a series of small holes around the perimeter of the desired shape, knocking out the piece and filing to shape. The earth screw hole is 4mm in diameter.

At the opposite end of the box, holes are required for the potentiometer and for a further earth screw. We used a countersunk screw here for the earth screw so that the end panel label would cover over the screw. In fact, we slightly 'over-countersunk' this hole to ensure the screwhead was flush with the case surface.

Insert the PCB into the case. Note that the leads for Q1 must be kinked outward a little so that the metal flange of the device is parallel to and in contact with the side of the case. Mark the mounting hole position for Q1.

TH1 also mounts on the side of the box adjacent to Q1; its attachment bracket is positioned so that the holes are vertical – the top hole about 7mm down from the top edge of the box.

Both the TH1 mounting screws and that for Q1 are 3mm countersunk. Countersunk screws allow the heatsink to mount flat to the surface on the side of the case without too much counter boring in the heatsink where these screws sit.



Another view of the opened case, including the back of the architrave GPO. Note the earthing of the case lid – we don't rely on the metal-to-metal contact. Also note that the circuit ground and the case earth are most definitely NOT connected together – the circuit ground in fact 'floats' at the mains voltage.

Resistor Colour Codes

	No.	Value	4-Band Code(1%)	5-Band Code (1%)
❑	4	1MΩ	brown black green brown	brown black black yellow brown
❑	1	220kΩ	red red yellow brown	red red black orange brown
❑	1	200kΩ	red black yellow brown	red black black orange brown
❑	1	22kΩ	red red orange brown	red red black red brown
❑	1	10kΩ	brown black orange brown	brown black black red brown
❑	2	5.1kΩ	green brown red brown	green brown black brown brown
❑	1	3.3kΩ	orange orange red brown	orange orange black brown brown
❑	3	1kΩ	brown black red brown	brown black black brown brown
❑	2	470Ω	yellow violet brown brown	yellow violet black black brown
❑	1	100Ω	brown black brown brown	brown black black black brown

Capacitor Codes

Value	μF value	IEC code	EIA code
100nF	0.1μF	100n	104
1nF	.001μF	1n0	102

The two 220nF, 250VAC 'X2' class will have values printed on them.

The 10μF and 100μF electrolytics can be replaced by surface-mount ceramic types (soldered to copper side of PCB).

Specifications

Rating	80W maximum, fused at 1A, 230VAC
Speed adjustment	Zero to maximum
Current limiting	235mA at low speed, up to 940mA at high speed
Temperature cut out	60°C (with 40°C cut in after 60°C cut out)

Note that you will find it easier to install TH1 if the M3 nuts are tack soldered to the thermostat mounting bracket. To do this, place the screws into the thermostat mounting bracket (when it is out of the case) and screw on the nuts. Solder the nuts in place by applying solder to the side of the nuts.

The aptly-named fan type heatsink is secured to the side of the case on the Q1 side, using two M4 screws that either tap into the side off the case or use nuts. The mounting holes are placed along the centre line of the heatsink. The lower hole should be positioned high enough so it does not foul the PCB, especially if using nuts. The heatsink is positioned with its lower edge at the same level as the bottom edge of the box.

The holes for Q1 and TH1 mounting must be countersunk; we actually over-countersunk them by about 0.5mm to ensure that the tops of the screws were actually lower than the surface of the case – this allowed the heatsink to make intimate contact with the case and therefore ensure maximum heat transfer (aided by a dollop of heatsink compound).

Holes are also required in the lid to secure the switched mains outlet and the earth terminal. We used a countersunk screw here for the earth screw so that the front panel label would cover the screw.

All holes must be de-burred on the inside of the case with a countersinking tool or larger drill to round off the sharp edge of the hole. This is especially so for Q1, where the edges must be rounded to prevent punch-through of the insulating washer. Run your finger over all holes to ensure there are no sharp edges – and also make sure no swarf is hiding in any of the box corners!

Panels

Artwork for the lid and end panels can be downloaded from the *EPE* website. Print them onto overhead projector film, photo paper or plain paper. We recommend overhead projector film – if you print in reverse, when it is placed on the box the printing will be against the case and protected by the film. The printouts can be cut to shape and adhered to the case with glue or silicone sealant.

Note that the countersunk earth screws for the lid and end panel need to be placed in position and temporarily held with a nut before placing the panels on.

Insert the PCB into the case by angling it so that the potentiometer is inserted into its hole first, then positioning the board onto the integral mounting lands inside the case. Secure the PCB to the case with the two 'supplied with the case' screws, plus the extra two M3.5 × 6mm screws.

Secure Q1 to the case with an M3 screw and nut with a mica insulating washer and insulating bush, as shown in the inset on the wiring diagram. Apply a smear of heatsink compound on all mating surfaces before assembly. We use a mica washer in preference to a silicone washer because the mica has a higher thermal conductivity (lower °C per Watt value) and the mounting screw can be tightened up more. This keeps the MOSFET cooler.

After mounting Q1, it is essential to check that the metal tab of the device is isolated from the case by measuring the resistance between the two with a multimeter. The meter should show a very high resistance measurement between the case and any of Q1's leads. Check your meter also reads close to zero ohms with a case-to-mounting-screw measurement. This will test if the multimeter is working and connected correctly.

The heatsink is attached using the two M4 screws, with a smear of heatsink compound between the mating surfaces.

Wiring

The complete wiring diagram is shown in Fig.4.

All mains wiring must be done using 250VAC, 7.5A mains-rated wire. You will need 200mm lengths of this wire in appropriate colours – brown (live), blue (neutral) and green/yellow (earth). The easiest way to get these (if you're not building from a kit) is to cut off a 200mm length from a spare piece of 230V mains flex, strip off the outer insulation and – *voila*!

The earthing details of the case are most important since D1 and the potentiometer are all at mains potential yet are attached to the case. If the insulating washer or the insulation of

Traditional (switched) fan speed controllers: how they work

The circuit at right shows a typical switched-type fan speed control. The fan motor has two windings, with one winding powered at a different phase to the other to provide a rotating field. To achieve this, the 'aux' winding is usually connected via a capacitor – in this case, 1.5µF.

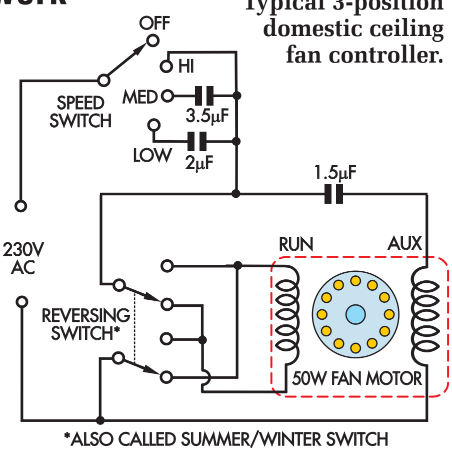
For speed control, this also uses capacitors (or sometimes inductors) to reduce applied voltage to the motor 'run' winding.

On the Hi setting, this winding receives full 230V AC power, so operates at maximum speed. When on the medium speed setting the run winding receives power via a 3.5µF capacitor in series and via a 2µF capacitor when switched to low speed.

At the 50Hz mains frequency, the 3.5µF capacitor has a reactance of 910Ω, so the motor runs quite a bit slower than on full power. A 2µF capacitor has a reactance of about 1.6kΩ, so the motor runs that much slower again. Note that lowering the capacitance increases the reactance (at that frequency).

Some ceiling fans also have a summer/winter switch, usually mounted on the fan itself, which simply swaps the connections to the run winding. This reverses the motor rotation, to push air in the opposite direction.

Typical 3-position domestic ceiling fan controller.



the potentiometer were to break down, the case would be live (ie, at 230VAC) if it was not properly earthed.

The potentiometer needs earthing since the screw thread does not reach far enough to the outside of the box for its nut to be screwed on to hold it to the case.

The potentiometer is earthed to the case by wrapping the earth wire around the location tab on the potentiometer and bending this down against the front of the pot. The earth wire is then soldered to this lug, ensuring there is sufficient heat for solder to flow onto the tab and wire – you may need to file or sand the lug first to remove any oxidation/passivation. Be certain that the solder joint on the tap is not a dry joint.

The case lid is also independently earthed, as shown.

The IEC connector must be wired using the correct wire colours – brown for the live, blue for the neutral and green/yellow striped wire for earth. Use insulated quick connectors for the mains wiring connection to the PCB. Wires to the IEC connector need to be insulated with heatshrink tubing covering all exposed metal terminals for the live and neutral wiring.

Solder two earth wires onto the earth pin on the IEC connector – one about 50mm long and the other about 150mm. These wires should loop through the hole in the earth terminal with each wrapped back on itself so the wires are essentially captive before soldering to the terminal. Make sure the earth terminal is heated sufficiently with the soldering iron so the solder wets and adheres properly to both earth terminal and wire. Again, be certain that it is not a dry solder joint. One end of the earth wire is crimped to the earth eyelet and the other to the earth eyelet on the lid and the GPO's earth terminal.

It is important to use one continuous earth wire length for the lid earth wire and GPO earth wire. Do this with just the insulation stripped back in the wire length to terminate into the crimp eyelet for the earth before running to the GPO's earth screw terminal.

The earth eyelets are secured with M4 screws, a star washer and nut. A second nut should be used as a locknut. As mentioned earlier, a countersunk screw is used for the earth on the lid and the potentiometer end panel – earth screws are placed before the labels are glued on. The IEC connector is secured with the M3 × 10mm countersunk screws, star washers and nuts. Similarly, the GPO is secured with M4 screws, star washers and nuts.

Finally, wires are secured using cable ties as shown.

Your speed controller is now complete – but don't forget to place four rubber feet on the bottom of the case if you want to avoid scratching surfaces underneath.

Testing

Check all of your wiring very carefully against the overlay and wiring diagram. Also check that the case, lid and potentiometer are connected to the earth pin of the power socket – use a multimeter on a low-ohms scale.

When you are satisfied that all is correct, screw the lid onto the case.

Note that while the case is supplied with a rubber seal that goes around a channel in the lid to ensure its IP65 rating, we elected not to use this, so heat from the case can transfer to the lid for maximum dissipation.

Do not be tempted to operate the fan speed controller without the lid in place and screwed in position.

The easiest way to test the circuit operation is to connect a fan. First, set VR1 fully anticlockwise, then plug a fan in, connect power and check that you can vary the speed with VR1. Note that the fan controller box will begin to run quite warm with extended use when driving the fan at lower than full speed. This temperature rise is normal.

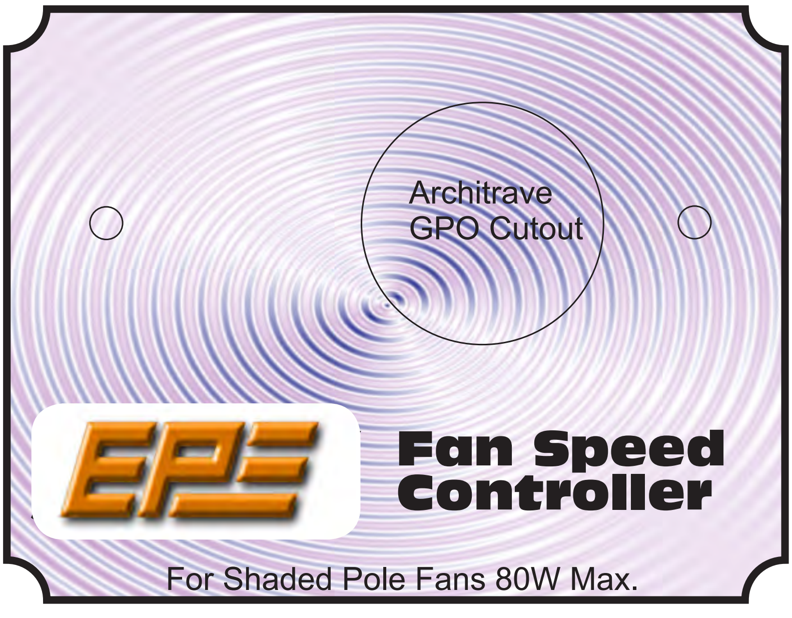
Troubleshooting the Fan Speed Controller

If the *Fan Speed Controller* does not work when you apply power, it's time to do some troubleshooting.

First, a reminder: all of the circuitry is at 230VAC mains potential and can be lethal. This includes any exposed metal parts on components except those that are tied to the earthed chassis of the case.

Do not touch any part of the circuit when it is plugged into a mains outlet. Always remove the IEC plug from its mains connector before touching or working on any part of the circuit. Before going any further, give your PCB another thorough check. Check for incorrectly placed components and for component orientation. Again, check solder joints.

Unless you have placed a component incorrectly or a solder joint is not properly made, there is very little that can go wrong with the circuit. It either works or it doesn't! So, if the *Controller* still doesn't work, check component placement and soldering once again.



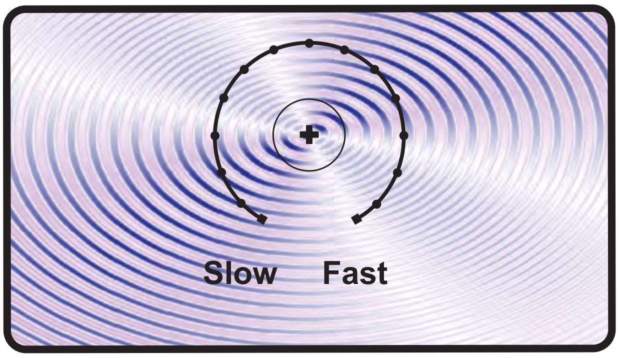


Fig.5: top-of-case and side-of-case panel artwork. This can also be downloaded from the *EPE* website and printed, in colour on thick paper or on overhead projector film.

Controlling ceiling fans

While this project was originally designed as a controller for free-standing fans – ie, those that plug into a mains outlet – there is no reason why it cannot be used for permanently installed ceiling fans. Of course, this would mean that the box would have to be mounted on a wall with wiring into the ceiling fan connections installed by a licensed electrician.

Any existing 'hard wired' switched-type controller could be left *in situ* – you'd simply leave it on its maximum setting and control the speed with this project.

There would obviously be no need for either the GPO on the case lid nor the IEC connector. Instead, wires would pass through cord-grip grommets or cable glands located in the side or base of the case.

You'd also need to fit a safety fuseholder in place of the one integrated with the IEC connector.

Wiring

Wiring details for direct connection are shown below. The 230VAC mains wires pass through grommets and the neutral connects directly to the PCB as shown.

The live is connected to a separate panel-mounted safety fuse holder (for the 1A fuse) that mounts on the case (or on the lid – ensure that it doesn't touch components underneath when the lid is screwed on).

The live and neutral outputs from the fan controller then connect to the existing fan speed controller at its 230VAC input. **The box must be earthed with earthing to the case, lid and pot body**.

Note that the speed control box needs to be mounted so there is access to the control knob and so the box can keep cool (ie, you couldn't mount it in a small wall cavity). The diagram below doesn't show the heatsink but it must be fitted, in exactly the same way as detailed earlier.

While we haven't confirmed it, we don't believe you could use this project and an electronic controller together – if you couldn't remove the electronic controller, you could simply bypass it.

And as a light dimmer?

This circuit will also make a fine incandescent light dimmer and, as we mentioned earlier, won't put lots of impulse noise onto your mains wiring to swamp AM radio reception.

So for fancy incandescent bulbs, spotlights, etc (up to 60W) it will be fine to use as is.

And if you are talking about a standard lamp that plugs into the power outlet, the unit can be constructed as detailed earlier, without changes. Permanent installation would require the wiring diagram below to be followed.

However, like old-style (phase-controlled) light dimmers, it is not suitable for CFLs nor any other lights (LEDs, for example) which have electronic controllers (remember that most LEDs these days have them either inbuilt or as part of the fixture).

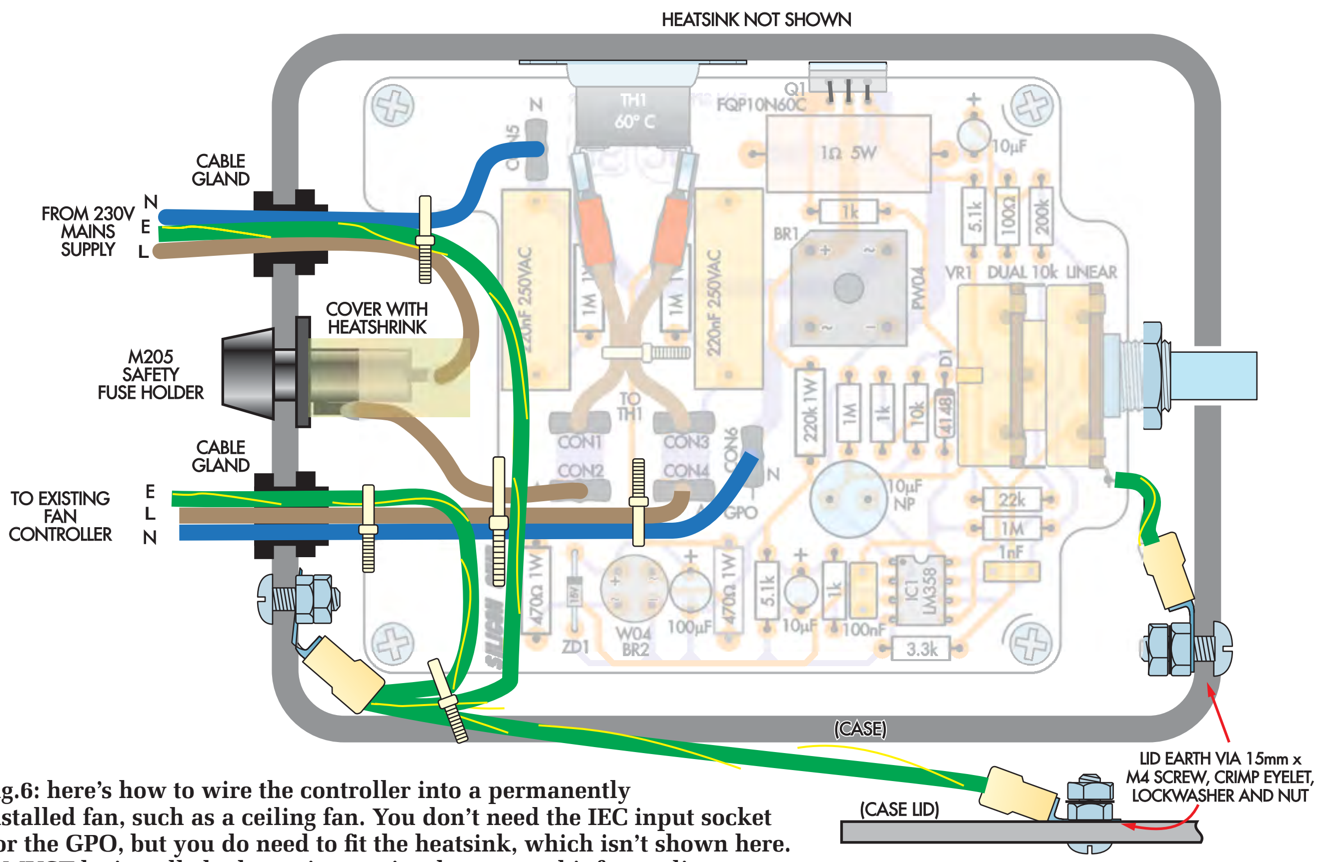
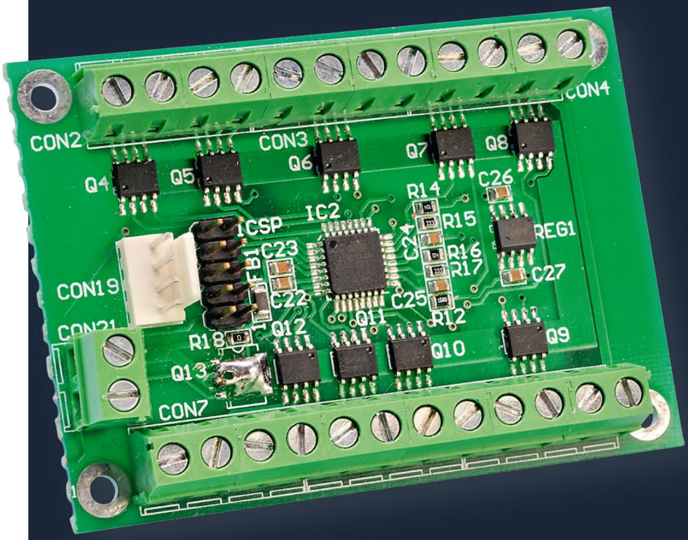
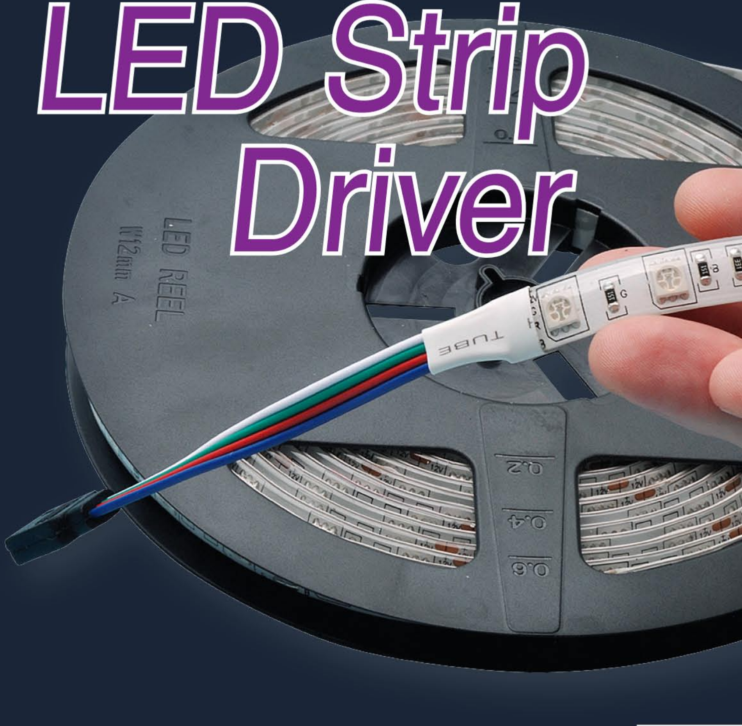


Fig.6: here's how to wire the controller into a permanently installed fan, such as a ceiling fan. You don't need the IEC input socket nor the GPO, but you do need to fit the heatsink, which isn't shown here. It MUST be installed where air can circulate around it for cooling.

Phantasmagorical RGB LED Strip Driver





This small module drives up to six RGB (red/green/blue) flexible LED strips to produce a rainbow of colours in multiple eye-catching patterns. Use it to decorate a Christmas tree, a shop window or anywhere else you want a bright, pulsating and flashing light show with many colours. It runs off a battery or a DC supply.

THIS PROJECT WAS DESIGNED to be used on a float in a street parade. I was helping a friend who was helping a friend to decorate the float and they wanted multiple flexible strings of LEDs, all constantly changing colours.

When I first heard about this, the plan was that they were going to build the electronics by hand, using through-hole components on Veroboard and point-to-point wiring – to drive around 30 RGB strips. I've built many prototypes this way and knew it was a dull and laborious process and the resulting boards can be quite delicate. So in order to head off the inevitable frustration I offered to design a 'proper' PCB.

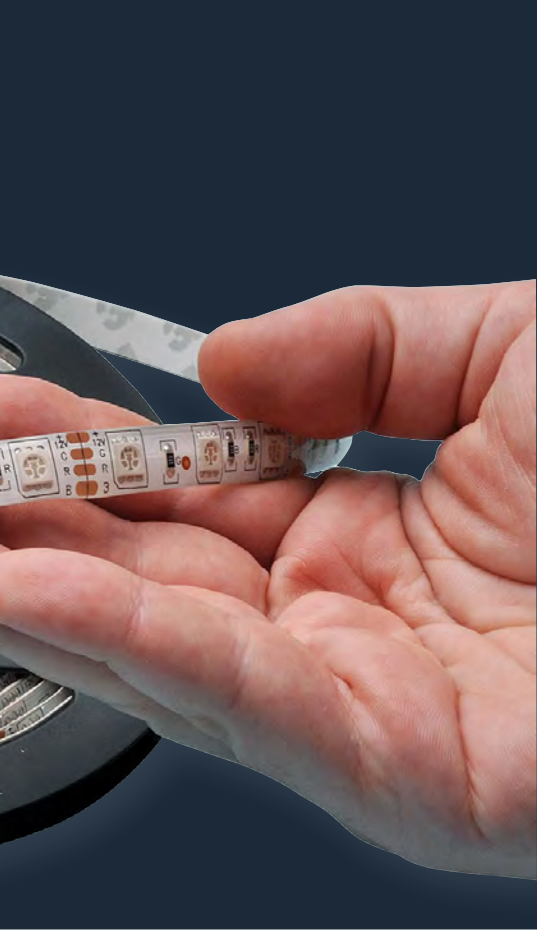
This was two weeks before the parade, so the design and assembly was a pretty quick affair. The boards were designed to be fast to build – I actually had to do them all in one evening after work, and managed to assemble the five boards in just four hours and deliver them to be programmed and wired up.

I didn't see them in action – but apparently they worked quite well, although by the end of the parade the batteries were pretty flat. We didn't have time to put in a low-battery cut-out feature, something which has been rectified in the final design. Obviously, this board is not limited to use on a float, so after some tweaking, we are publishing it for general use.

Design

RGB LED strips can be purchased on 5m reels, made up of 100 joined

By NICHOLAS VINEN



sections, each 50mm long. They are also available in shorter lengths. Fig.1 shows a typical arrangement.

These components are mounted on a long, thin flexible PCB with a plastic cover over the top and in our case, with an adhesive backing. Power consumption is around 7.5W/m (375mW per section) at 12V, with all LEDs at full brightness. We measured 920mA for blue, 1150mA for red and 1040mA for green on a 5m strip.

Our reels were supplied with mating 4-pin plugs at either end (2.54mm pin spacing), so they can be combined into longer lengths if required. If you cut the strip up into shorter lengths, this exposes a set of four pads on either side of the cut, to which a similar cable can be soldered. We got ours from an Internet seller, but very similar products are widely available.

To control these strips to get any colour we want, we apply 12V to the anode terminal and then vary either the resistance or (in this case) PWM duty cycle between the cathodes and ground, thus varying the red, green and blue component brightness. These colours combine – so, for example, if all three are driven at a similar level, the resulting light looks (more or less) white. Or if red and blue are driven, but green is not, the result is mauve.

Now since we ran our strips off a battery, the supply voltage wasn't constant (this will also be true if the power source is unregulated 12V DC from mains). In fact, the Li-Po batteries we used were 4-cell packs with a full charge voltage of 4 × 4.2V = 16.8V and a flat voltage of 4 × 3V = 12V. An unregulated mains-powered 12V DC supply would have a similar voltage range, but regulated supplies are more common at the high currents required. A discarded PC power supply would be eminently suitable.

If we simply ignored the varying battery voltage, the LED strips would dim over time as the batteries discharged and we would also risk burning the strips out when the battery is fully charged and the supply is significantly higher than the 12V that the strips are designed to be driven with.

One way to avoid this would be to regulate the supply to a constant 12V, but a much easier method is to figure out how the brightness of each colour varies as the supply goes above 12V, then monitor the supply voltage and reduce the duty cycle to compensate, giving constant brightness. This is a very efficient way to do it because very little power is lost and it also minimises the component count.

Since we only need to switch the LED cathodes, this makes the circuit design easy. For each colour of each strip we just need one low-side switch, and an N-channel MOSFET does the job. These are available in dual SMD packages, which are quite compact and easy to solder, with suitable voltage and current ratings and an on-resistance figure of around 10mΩ. So for each FET handling 1A, the dissipation is only 10mW.

To further simplify the circuit, the MOSFET gates can be driven directly from the outputs of a microcontroller and this is much easier for low-side switching than high-side switching. But we do have to be a little careful, since microcontroller outputs can provide relatively little current (typically ~40mA DC and 100mA peak) and we also need to make sure we don't exceed the micro's ratings.

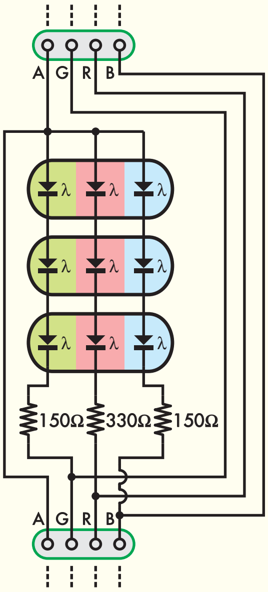


Fig.1: the circuit diagram of a section of typical RGB LED strip. This is repeated every 50mm, with the connectors at top and bottom joined end-to-end. The strip can be cut into any number of whole sections (up to the maximum of 100 supplied on the reel) and can be driven from either end. The more sections you drive, the more current it draws – see text for details.

The switching time of a micro output, driving the small capacitance of the type of MOSFET we're using is quite fast at around 100ns, so that isn't really an issue. But when driving 6 × 3 = 18 MOSFETs from a single micro, the instantaneous current is a concern should they all switch simultaneously. The micro we're using has an absolute maximum rating of ±40mA (DC) per output pin and 400mA for the whole device.

Examination of the I/O pin source/sink current vs output voltage graphs suggests that the output transistors have an on-resistance of around 100Ω. So if eight outputs are switched simultaneously (the maximum possible with an 8-bit micro) to discharge MOSFET gates at 5V, the total current at that instant would be (5V ÷ 100Ω) × 8 = 400mA. That's just equal to the rating, but it's also only for a brief period; as the gates discharge, the sink current rapidly drops. So we don't see any problems with this arrangement.

Battery protection

We also need to consider the health of the battery. A lead-acid battery could be used and these can be discharged

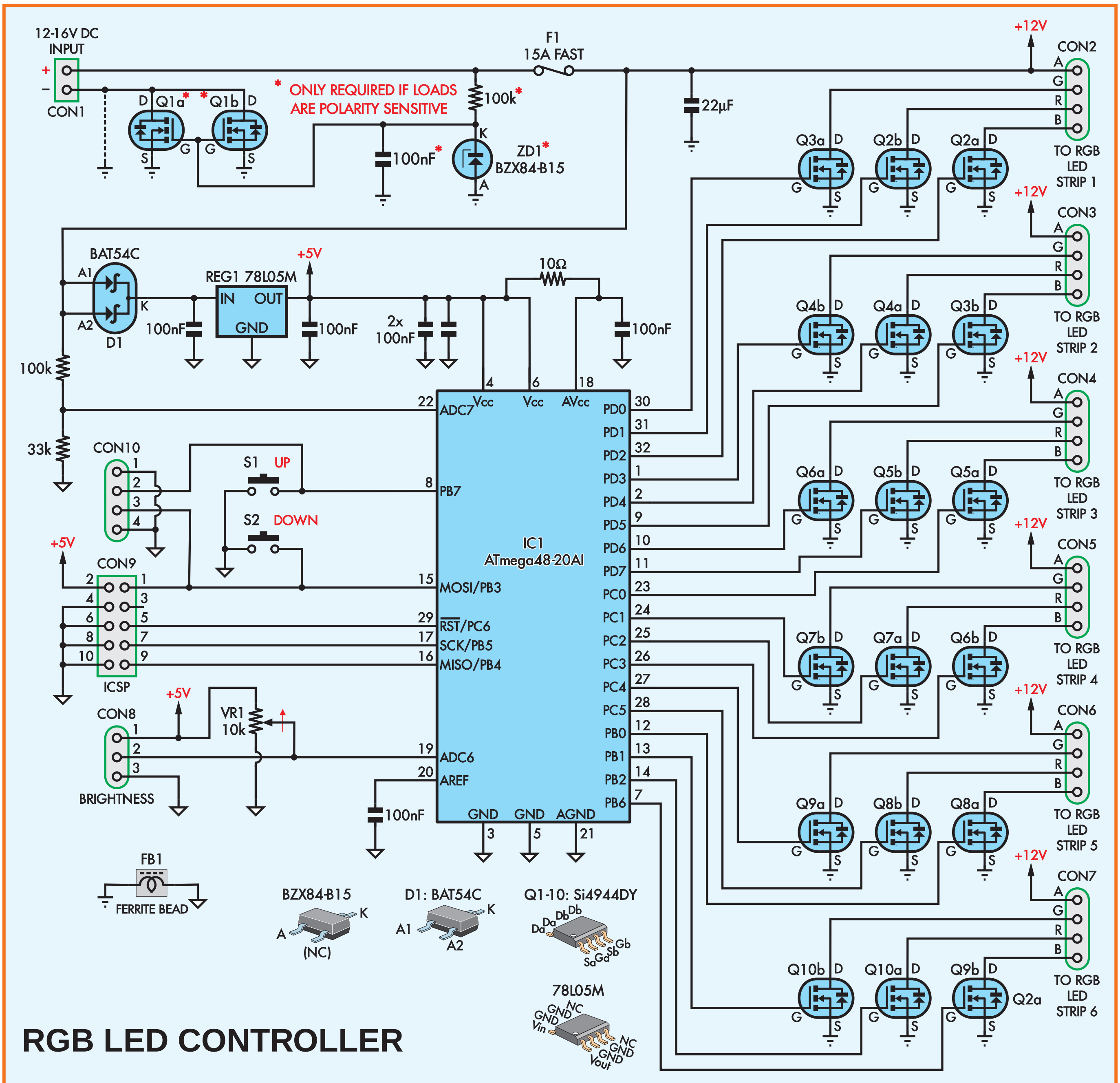


Fig.2: the complete circuit diagram of our 6-strip RGB LED driver. It's a simple affair with microcontroller IC1 driving the gates of 18 MOSFETs directly to control the cathodes for three strings of LEDs in each of six connected strips. REG1 derives power for the micro from the nominal 12V supply, while S1 and S2 allow the pattern to be changed and VR1 varies the overall LED brightness.

to about 11.5V before being damaged, but by then the battery will be well and truly flat and the LED strips will be noticeably dimmer.

Li-Po batteries should not be discharged below about 3V per cell, ie, 12V for a 4-cell pack, or else they can be destroyed. So to be safe, the unit should stop drawing current once the battery voltage drops much below 12V.

We're already monitoring the supply to provide LED PWM duty cycle compensation, so it's simply a matter of programming the micro to turn off all the outputs and go to sleep if the battery voltage drops too low. It can then periodically wake up to check the voltage and if it recovers sufficiently (eg, the battery is under charge), it can then go back to normal operation.

In sleep mode, the only part of the circuit drawing any significant current is the 78L05M regulator at about 3mA. With the large battery required for this project, that will give you several days to disconnect the unit and recharge the battery before it goes totally flat. This time could be extended dramatically by replacing the regulator with a lower quiescent current type, but in most cases this should not be necessary (the micro draws <1µA in sleep mode).

Circuit description

Fig.2 shows the full circuit. The LED strips are wired to 4-way terminal blocks

Parts List

1 double-sided PCB, available from the *EPE PCB Service,* coded 16105141, 82 × 55mm
1-6 RGB LEDs or LED strips
1 12V DC power supply or 12V battery
13 2-way PCB-mount terminal blocks, 5.08mm spacing, rated at 15A+ (CON1-CON7)
1 15A SMD fuse, 3216 or 6432 size (1206/2512 imperial) (F1) (element14 2135886, Digi-Key 507-1059-1-ND)**
1 mini horizontal 10kΩ trimpot (VR1) (optional) OR
1 3-pin header (CON8) plus external pot and wiring (optional)
1 5 × 2 pin header (CON9) (not required with pre-programmed microcontroller)
2 PCB-mount tactile buttons (S1,S2) OR
1 4-way pin header (CON10) plus external buttons and wiring
1 SMD ferrite bead, 3216 size (1206 imperial) (element14, RS, Digi-Key)

Semiconductors
1 Atmel ATmega48-20AI or -20AU 8-bit 4KB microcontroller programmed with 1610514A.HEX (IC1): element14 Cat 9171312, Digi-Key ATMEGA48-20AU-ND***
1 78L05M SMD 5V 100mA regulator (REG1) *
9 Si4944DY SMD dual N-channel MOSFETs or equivalent (Q2-Q10) *
1 BAT54C dual common-cathode Schottky diode (D1)*

Capacitors
1 22µF 25V SMD ceramic, 3216 size (1206 imperial) (element14 2354129, Digi-Key 1276-3047-1-ND)
7 100nF 50V SMD ceramic, X7R, 1608 or 2012 size (0603/0805 imperial) (element14 1301790/1301894, Digi-Key 1276-1180-1-ND/311-1344-1-ND)

Resistors (all SMD 1608 or 2012 size [0603/0805 imperial])
1 100kΩ* 1%   1 10Ω*
1 33kΩ* 1%

* Available from element14, RS and Digi-Key
** Spare SMD fuses are handy
***Programmed Atmel ATmega micros available from: **www.siliconchip.com.au/Shop/9**

CON2-CON7 and MOSFETs Q2a-Q10b switch the cathodes, with the anodes all connected together to the (nominal) 12V supply. This supply comes via input connector CON1 and passes through a 15A PCB-mount SMD fuse, which we put in as the last-ditch protection against a serious fault, such as a shorted output (Li-Po batteries don't like to be shorted out). A 22µF capacitor smooths this supply and reduces its impedance.

The micro we've used is an ATmega48 in a 44-pin SMD package. We chose this because it's easy to program and as described above, has good output drive capability for switching the MOSFET gates. Its 5V supply is derived from the fused 12V rail via reverse-polarity-protection Schottky diode D1 and REG1. D1's two internal diodes are paralleled for lower losses and higher current capability.

The micro has a 100nF bypass capacitor for each of its V_{CC}/AV_{CC} (analogue supply) inputs. AV_{CC} is smoothed by a low-pass filter formed by a 10Ω resistor in combination with its 100nF bypass capacitor.

The 12V supply is monitored using a 100kΩ/33kΩ resistive divider from that rail to ADC input 7 (pin 22). This 4:1 divider gives a voltage at pin 22 of 2.875 – 4.25V (11.5 – 17V supply) which is measured relative to the 5V rail. A 100nF capacitor from the A_{REF} pin (pin 20) to ground filters switching noise from the reference voltage, which is derived from AV_{CC}.

The microcontroller can be programmed via a standard 10-pin Atmel AVR in-circuit serial programming (ICSP) header (CON9). However, if you use a pre-programmed micro (from: **www.siliconchip.com.au/Shop/9**) then CON9 could be omitted.

The original design had a fixed LED display pattern, but we decided to revise it to give multiple patterns, hence the addition of pushbutton switches S1 and S2. These are connected to input pins PB3 and PB7 of IC1, which have internal pull-ups enabled. S2 shares a line with the programming header, which is fine as long as you don't press it during programming.

CON10 allows off-board buttons to be used instead of S1/S2 if desired. Trimpot VR1 gives overall LED brightness control or an off-board pot can be wired to CON8 which is fitted in place of VR1. You can also simply solder a wire link between pins 1 and 2 of CON8 so that the LEDs run at full brightness all the time.

The ground connection for switching MOSFETs Q2-Q10 is kept separate from the ground for the rest of the circuit, hence the use of two different symbols. These two grounds are joined at a single point by ferrite bead FB1, which reduces the coupling of switching noise into the microcontroller's ground, thus reducing errors in its ADC readings.

FB1 is shown in the botton left corner of the circuit, connecting MOSFET ground to the input supply ground.

Finally, note that we show components to protect the load from reversed polarity on the input connector. These are Q1, ZD1, and a 100nF capacitor and 100kΩ resistor. However, the LED strips are unlikely to be damaged by reverse polarity, so they probably do not need to be installed. (**A track on the board (shown dashed) connects the ground return directly to CON1 and must be cut if Q1 is to be fitted.**)

We've left provision for these components on the PCB, in case a different type of load is connected which is polarity sensitive.

Note that fuse F1 is a surface-mounting component and if it blows you will have to de-solder it and solder another in its place. However, with some care in wiring the unit up and ensuring that it's used within its ratings, there's no reason for it to blow. If you aren't planning to use the full-current capabilities of the device, eg, your load will never exceed 10A, it's a good idea to fit a fuse with a lower rating (but higher than the expected maximum load current). You could also use an inline fuse from the battery, which would be easier to replace.

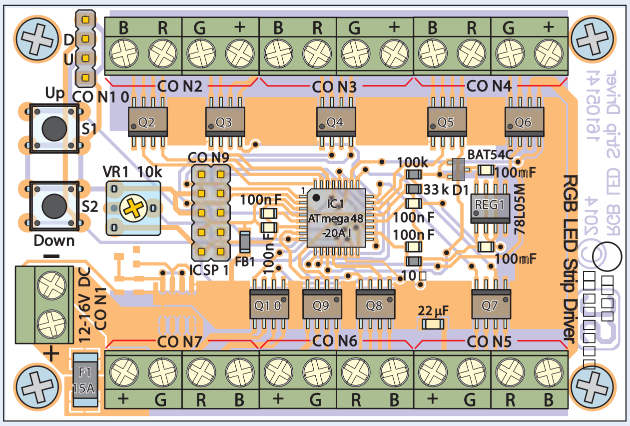
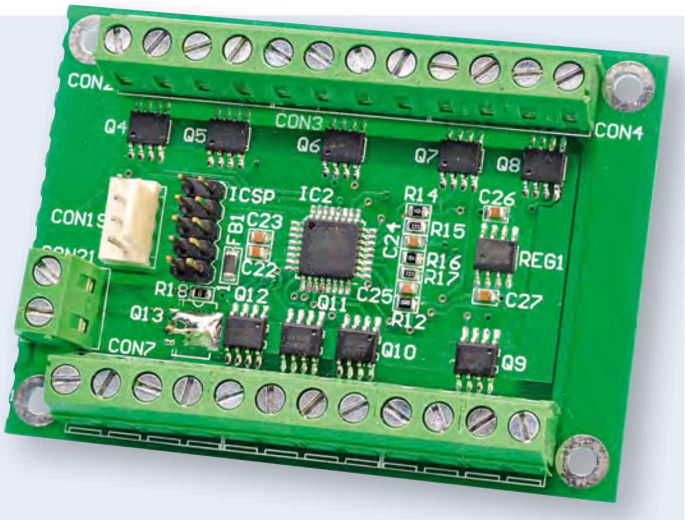


Fig.3: the PCB is quite compact and is fitted mostly with surface-mounting components, the exceptions being the connectors, pushbuttons S1 and S2 and trimpot VR1. S1, S2 and VR1 can also be mounted off-board to give external controls or left out entirely if their functions are not needed (VR1 must be linked out in this case).



Above: this photo shows a completed prototype PCB assembly. Note that the final version shown in Fig.3 has a few changes, including the addition of trimpot VR1, pushbutton switches S1 and S2 and SMD fuse F1.

Software

Since this chip only has a handful of PWM channels, we have to use the outputs as general purpose I/Os and arrange the software to provide PWM by constantly updating these output states. They have been arranged to make it simple for the software by wiring up the MOSFET gates to sequentially numbered pins.

The micro runs at 8MHz with one of its internal timers configured to divide-by-128 to give 62.5kHz. It then divides this by 256 brightness levels to get 244Hz PWM operation. The main loop continuously calculates the next state of each output as an RGB value from 0-255 (ie, from off to maximum brightness) and then computes the timing for switching the MOSFETs off and on to achieve this. The timer interrupt is then set to trigger a subroutine at the right times to turn the outputs on/off to achieve this pattern.

This repeats indefinitely. It periodically stops to check the position of VR1 and whether S1 and/or S2 have been pressed. If so, it switches patterns.

PCB assembly

The PCB assembly is relatively straightforward, with no particularly difficult-to-solder parts, but some care does need to be taken to ensure the SMD solder joints are properly formed and there are no bridges. Start with the SMD ICs and MOSFETs, then follow with the passive SMDs and finish up with the through-hole parts.

IC1 is probably the best one to do first. This is installed by positioning it on the board with the correct orientation, placing some solder on one of its pads and heating that pad while sliding the IC into place. You should then check its alignment. Make sure all the pins are properly centred on the pads and then solder the diagonally opposite pin. Make a final check that the orientation is correct, then solder the rest of the pins.

It's possible to solder each of IC1's pins individually with a fine-tipped soldering iron, but it is not necessary to do so. You can place the tip of the iron between a pair of pins and flow solder onto both, then clean it up later using solder wick. You could also use a mini-wave/hoof tip or one of various other methods such as hot-air or oven reflow.

It's a good idea to use flux paste, both to aid the initial soldering and in combination with solder wick if cleaning up any bridges is necessary. When finished, clean off any flux residue with a good solvent, then inspect the joints carefully under a magnifying glass with good illumination. Check that they have all formed good fillets between the IC pins and the PCB.

Next, you can fit MOSFETs Q2-Q10 and regulator REG1. Pay close attention to the pin 1 marking which may be a dot or bevelled edge and make sure the 78L05M goes in the right place. The pin spacing on these parts is larger than IC1, so it's realistic to solder the pins individually, although the techniques mentioned above remain valid. As with IC1, a careful inspection of the joints is most important.

Now fit D1 using a similar approach; you certainly can solder its pins individually. Then follow with the passives (resistors and capacitors) but remember to wait a few seconds after sliding the part into position before soldering the opposite side so that the first joint has had time to cool.

One way to check whether these components have been soldered properly is to heat one end and apply gentle pressure on the part with the soldering iron; if the opposite joint is bad, it will slide out of position and you will have to remove/re-solder it.

Assuming that the joint is OK, let it cool and then check the other using a similar method. However, after doing this you should inspect the joints and re-flow them if they look crystalline or lumpy.

Solder fuse F1 in place, then move on to the through-hole parts, starting with S1 and S2, or alternatively CON10, which is wired up to external buttons later. If you leave these parts off altogether, the unit will then be permanently set to pattern cycle mode.

Before fitting the terminal blocks, gang them up into two sets of six, using the integral slots and tabs. That done, make sure they are pushed down fully

Features and Specifications

Outputs: 6 × 3-channel 12V RGB LED strip drivers (common anode), up to 5A each strip (15A total maximum)

Input: 12-17V DC at up to 15A from battery (lead-acid, Li-Ion, Li-Po) or mains supply

Patterns: 10 different patterns plus auto-cycle mode which changes pattern periodically

Protection: fuse, reverse-polarity protection, battery over-discharge protection

Other features: constant brightness, optional brightness control

Battery cut-out: ~11.5V with 0.5V hysteresis

PWM frequency: ~250Hz

onto the PCB with their wire entry holes facing outwards before soldering all the pins. Then fit either VR1, a 3-pin header in its place or a wire link between the two lower pads. Finish off by soldering CON9, but note that it isn't necessary if you're using a pre-programmed microcontroller.

Programming

If using a blank micro, now is a good time to program it. First, connect a 12V supply (current-limited, if possible) to CON1 and check that there is 5V between pins 2 and 4 of CON9. You can then connect an AVR ICSP tool and upload the HEX file, which can be downloaded from the *EPE* website.

You will also need to set the 'fuse bits'. An unfortunate aspect of programming AVRs is that these are not included in the HEX file and there is no consistent way of referring to them. There are two bytes to set. Set the fuse high byte to 'DC' hex and the fuse low byte to 'C2' hex.

Depending on your programmer, you may not be able to set these as hex values so instead, for the high byte, set BODLEVEL to '100' (4.3V) and leave the rest of the settings at their defaults, ie, RSTDISBL = 1 (off), DWEN = 1 (off), SPIEN = 0 (on), WDTON = 1 (off) and EESAVE = 1 (off).

For the low byte, set CKSEL = 0010 (Calibrated Internal Oscillator), with CKDIV8 = 1 (off) and SUT = 00 (fast rising power). Leave CKOUT at its default value: CKOUT=1 (off). This sets operation to 8MHz, as the software expects.

Testing

There isn't much to test; check that the 5V supply is correct as described above and that the current draw is reasonable (<30mA), then connect a proper 12V supply and an LED strip to one of the outputs and power it back up. You should see the LEDs light up and the colour change over time. If so, you can then switch off and connect strips to the remaining outputs, switch back on and check that they are all operating and displaying the full range of colours.

Press S1 and S2 to see that the pattern changes and if VR1 is fitted, adjust it and check that it controls the brightness. Note that if you are using an off-board pot, this will need to be wired up for testing or else the results will be unpredictable (but no damage should occur).

Using it

Pressing S1 cycles to the next pattern and pressing S2 switches to the previous pattern. Initially, the unit starts with pattern 1, then after a minute or so switches to pattern 2 and eventually after pattern 10, it goes back to the first one. This cycle repeats 'forever', but it is cancelled by pressing either S1 or S2 after which it will remain on that same pattern. To switch back to auto-cycling mode, press S1 and S2 simultaneously.

VR1 adjusts the maximum duty cycle but note that the duty cycle is also automatically reduced as the supply voltage rises above 12V to give even brightness regardless of battery voltage (down to a minimum 12V). Note also that should the battery voltage drop below about 11.5V (including wiring drops), the unit will shut down until it rises above 12V or so.

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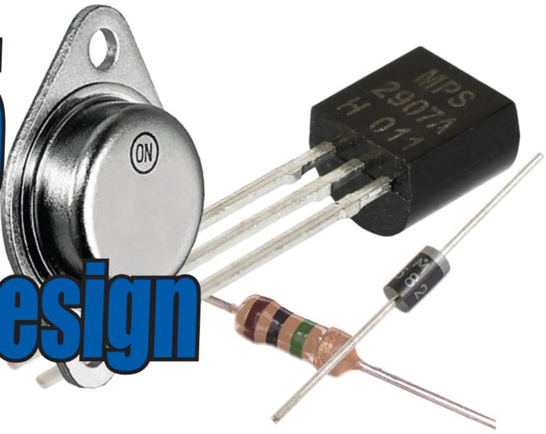
Teach-In 2015

Discrete Linear Circuit Design

Part 4: Distortion



by Mike and Richard Tooley

Welcome to *Teach-In 2015* – This series is aimed at anyone wishing to develop a detailed understanding of linear discrete semiconductor devices and how they are used in a diverse range of circuits. We hope you will join us on this exciting voyage of discovery!

Each part of our *Teach-In 2015* series is devoted to a different aspect of discrete linear circuit design, such as modelling and simulation, measurement and testing, noise and distortion. In last month's instalment, *Discover* introduced you to emitter followers and complementary output stages, while *Knowledge Base* explained the various classes of operation used for linear amplifiers. Our practical feature, *Get Real*, described the design and construction of a simple headphone amplifier that can be used to boost the output of MP3 players or similar portable music playing devices. We also showed how decibels (dB) are used to specify a variety of parameters in electronics, including gain and sound pressure level (SPL).

Introduction

In this month's *Teach-In 2015*, the *Get Real* section will explain some tests and measurements on the simple headphone amplifier that we described last month. *Knowledge Base* will introduce you to something that we need to avoid in linear circuit design – noise and distortion – while *Discover* will describe a simple but very effective method of checking and assessing the frequency response of an amplifier based on square wave testing.

Knowledge base: noise and distortion

At first sight, it might seem a little odd to be concerned with something that we don't actually want, but noise and distortion is a problem that we do need to get to grips with if we are concerned with reproducing a signal faithfully. Put simply, noise and distortion is anything introduced by our circuit that wasn't present in the input signal or anything that's removed from the original signal. The only quantity that can change is the amplitude of the signal – the waveform must be faithfully reproduced.

Just how much noise and distortion we can tolerate depends on the particular application that we are dealing with. A telephone conversation, for example, might be perfectly intelligible when a significant amount of noise and distortion is present, but the same amount of signal degradation would not be tolerable when listening to music from a good quality audio system. The human ear can detect quite small levels of noise and distortion, but the crucial factor is the level of noise and distortion when compared with the level of the wanted signal.

Types of distortion

There are several different forms of distortion, and all of them may, to a greater or lesser extent, be present at the same time. If an amplifier has a perfectly linear transfer characteristic and a perfectly flat frequency response it will not produce any distortion (it might, however, be susceptible to noise and interference, as we explain later). Conversely, if the frequency response and/or transfer characteristic is imperfect, then this will result in the production of distortion but this may, or may not, be a problem, depending upon the severity of the non-linearity and the degree of aberration in the frequency response. Let's first consider the effect of non-linearity in the transfer characteristic. This is often referred to as 'non-linear distortion', but before we delve a little deeper it is important to know about the fundamental and harmonic components that make up a waveform.

Fundamentals and harmonics

The sinewave is the most fundamental of all wave shapes, and all other waveforms can be synthesised from sinusoidal components. To specify a sinewave, we need to consider just three things: amplitude, frequency and phase. Since no components are present at any other frequency (other than that of the fundamental sinewave) a sinusoidal wave may be referred to as a *pure tone*. Adding together sinewaves of the correct amplitude, frequency, and phase can reproduce all other waveforms.

An integer multiple of a *fundamental* frequency is known as a *harmonic*. In addition, we often specify the order of the harmonic (second, third, and so on). Thus, the *second harmonic* has twice the frequency of the fundamental, the t*hird harmonic* has three times the frequency of the fundamental, and so on. Consider, for example, a fundamental signal at 1kHz. The second harmonic would have

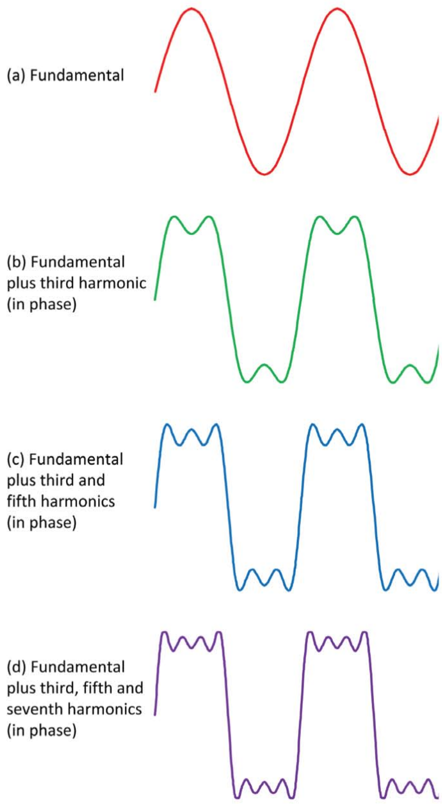


Fig.4.1 *Effect of adding odd harmonic components to a fundamental sinewave*

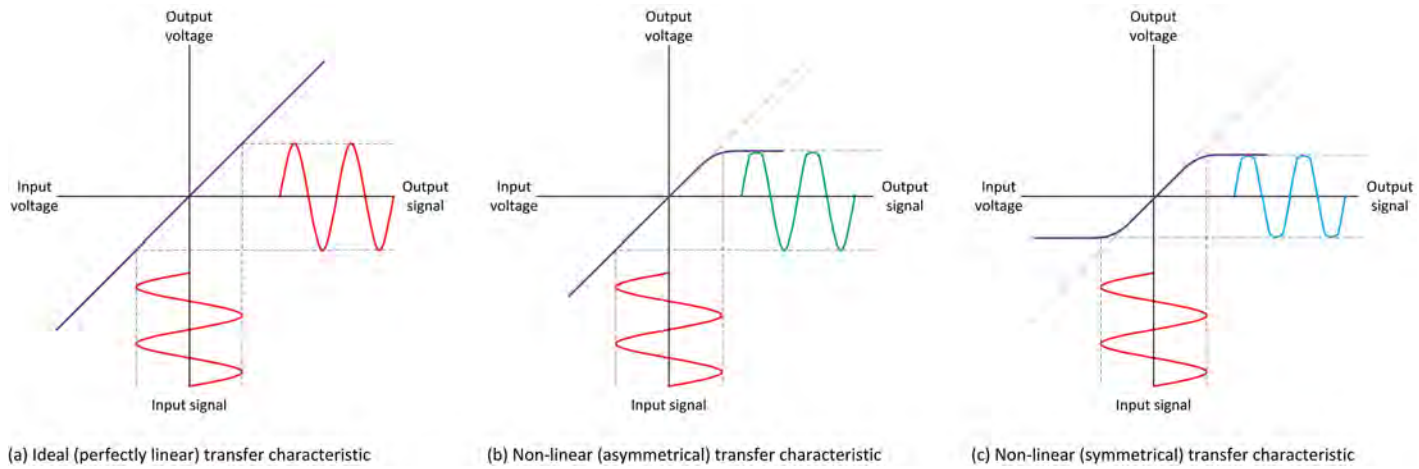


Fig.4.2 Effect of a non-linear transfer characteristic on the quality of an output waveform

a frequency of 2kHz, the third harmonic a frequency of 3kHz, and the fourth harmonic a frequency of 4kHz. Note that, in musical terms, the relationship between notes that are one *octave* apart is simply that the two frequencies have a ratio of 2:1 (in other words, the higher frequency is double the lower frequency).

Square waves

Square waves are comprised of a fundamental component together with an infinite series of odd harmonics (×3, ×5, ×7, ×9...). The amplitudes of these harmonic components decay with their harmonic order and they are all in phase with the fundamental. Thus, the third harmonic has an amplitude that is one third of the fundamental. The fifth harmonic has an amplitude that is one fifth of the fundamental, and so on.

The effect of adding several in-phase low-order harmonic components to a fundamental sinewave is illustrated in Fig.4.1. Notice how the resulting waveform becomes increasingly square as the higher order harmonics are added. Conversely, if these high order harmonics are not present (or if their amplitude is reduced) the waveform will become less square. This phenomenon leads us to a rather neat way of quickly assessing the frequency response of an amplifier when a reasonably accurate square wave is applied to its input. Later in this instalment we will explain how this is done.

Non-linear distortion

The two most common forms of non-linear distortion are harmonic distortion (often specified in terms of the total amount of harmonic distortion present, THD) and intermodulation distortion, IMD. In both of these forms of distortion, extra frequency components are added to a signal due to the non-linearity of an amplifier's transfer characteristic.

Harmonic distortion can be expressed in decibels or as a percentage. Even-order harmonic distortion is generally caused by an asymmetric transfer characteristic, whereas odd-order harmonic distortion is caused by a symmetric non-linearity. Fig.4.2 shows how this distortion arises. In Fig.4.2(a) we have shown a perfectly linear transfer characteristic (output plotted against input). In this condition, the output waveform will be identical to that present at the input.

Fig.4.2(b) shows what will happen when the transfer characteristic is asymmetric. In this case, the characteristic flattens-off beyond a particular positive going input level. The result is that the output signal becomes prematurely truncated or *clipped*. A similar effect would be produced if the negative-going part of the transfer characteristic had become flattened while the positive-going part remained linear. However, in this case the negative edge of the output signal would be clipped. The resulting harmonics produced will include the second, fourth and sixth, and so on, as shown in Fig.4.3(b). None of these components are present in the undistorted pure sinewave signal shown in Fig.4.3(a).

Fig.4.2(c) shows symmetrical non-linearity in the transfer characteristic. In this case, both positive and negative edges of the output waveform have become clipped. Notice also that the output amplitude has become reduced when compared with the undistorted output waveform shown in Fig.4.2(a). The resulting harmonics produced will include the third, fifth, seventh, and so on, as shown in Fig.4.3(c). Once again, none of these components are present in the undistorted pure tone shown in Fig.4.3(a).

It is usually accepted that odd-order harmonic components sound a lot worse than even-order harmonics. In a well-designed system, higher order harmonics

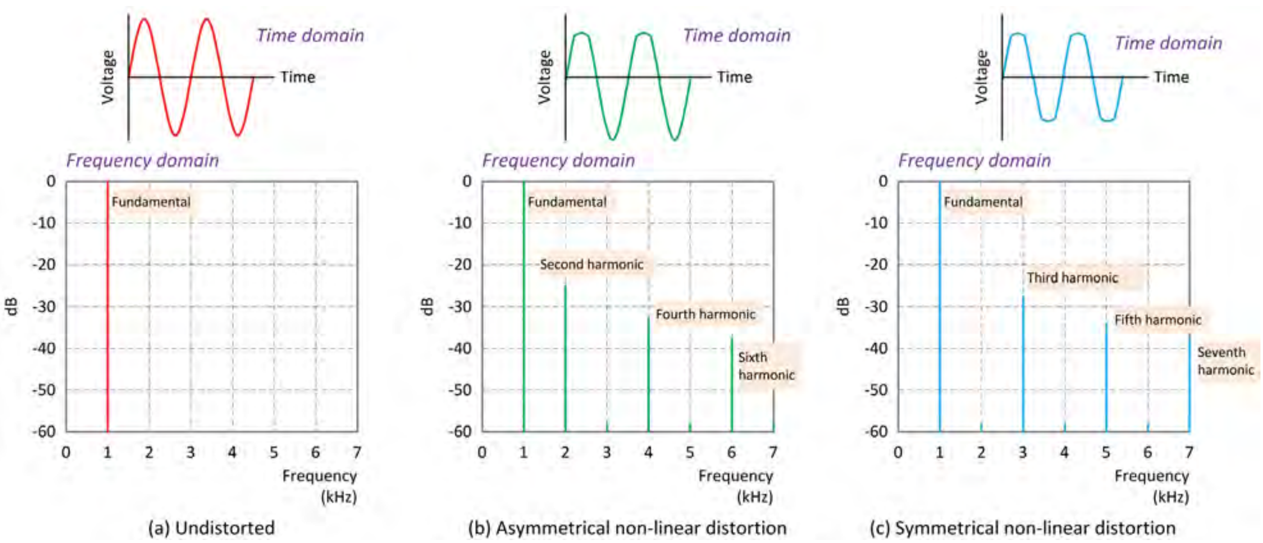


Fig.4.3 Harmonic analysis of the distorted and undistorted signals in Fig.4.2

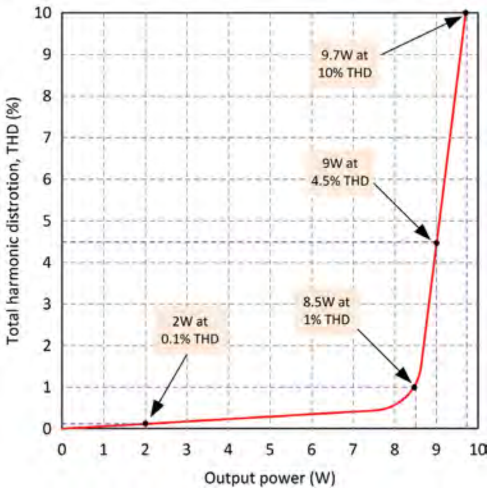


Fig.4.4 Overdriving a power amplifier results in a rapid rise in total harmonic distortion

should be significantly lower than the second and third-order harmonics. Note that harmonic distortion is non-linear with level and an increase in the amplitude of a test signal will usually result in a significantly greater increase in the level of harmonic distortion. This condition occurs whenever an amplifier is being *overdriven*. For example, an amplifier designed for a maximum input signal amplitude of 100mV peak-peak will be overdriven when a signal of 200mV is applied. The problem will increase in severity with a further increase in input signal to the point that the distortion produced will very quickly reach an unacceptable level. Fig.4.4 shows the effect of overdriving a power amplifier on the amount of total harmonic distortion generated. Note the rapid increase in THD when the rated output power (8W) is exceeded.

How much is bad?

This then begs the question of how much distortion you can actually tolerate. An acceptable level of distortion in a low-cost portable radio might be as much as 5%, whereas, in a high-quality audio system, distortion of 1% might be considered excessive.

Purists and trained audiophiles might disagree, but it is extremely difficult – if not impossible for most people – to discern the presence of total harmonic distortion (THD) when it is less than about 1%. Most well-designed amplifiers produce a THD level that is very much less than this (our simple headphone amplifier measured less than 0.3% THD when delivering 1mW of power to a 600Ω load).

Of significance (and often forgotten when people make claims for the quality of their audio systems) is the distortion that is introduced by the output transducer (ie, loudspeaker or headphones). In practice, this rarely quoted specification might be between 3% and 5% THD, and since this is invariably much greater than the THD produced by the electronics, any small amount of distortion in the amplifier will simply pale into insignificance.

It is also worth noting that at very low frequencies the THD introduced by a loudspeaker can be very significant (5 to 10% is not unusual) but this might not be such a problem as it first appears because research suggests that human listeners are very much less able to discern distortion when it occurs at low frequencies than when it arises at higher frequencies. Thus, 20% distortion at 50Hz might actually be less of a worry than 2% at 5kHz!

Note also that, when a loudspeaker is producing 3% THD, there will be no discernible difference when the transducer is used in conjunction with amplifiers having THD ratings of 0.03% and 0.0003%. In other words, there is little need for an amplifier to have a THD of a mere fraction of 1% when, in the real world, it will be used to drive a loudspeaker producing significantly more distortion. The weak link in the chain is the loudspeaker, not the amplifier.

How the THD is measured is also important. Measuring THD with a signal having a level just below the threshold of clipping will be instrumental in hiding other forms of distortion, notably cross-over distortion, noise and hum which all disproportionately affect lower signal voltages. In most situations a THD measurement performed at 10 dB below rated output power (eg, measured at 1W RMS output for an amplifier rated at 10W RMS output) would be a much better indicator of sound quality.

Distortion due to aberrations in frequency response

Note that an amplifier will also introduce distortion if it does not have an ideal frequency response. It can be a sobering experience to apply a square wave to an amplifier and find that the output waveform is not very square. The reason for this is that, for a square wave to be perfectly reproduced, an amplifier would need to have a perfect frequency response. This, of course, is never actually the case.

The reason for this is simply that a square wave comprises a fundamental frequency component (a sinewave) together with an infinite number of odd harmonic components (×3, ×5, ×7...). To reproduce this waveform faithfully would require an amplifier with a frequency response extending from DC (zero frequency) to infinity!

Noise

What do we mean by 'noise'? Put simply, noise is a random fluctuation introduced by a component or circuit that appears superimposed on a wanted signal. When attempting to hold a conversation with someone in a crowded room, all of the other conversations present become noise. They detract from the conversation that we want to hear and impair the quality and intelligibility of it.

Unfortunately, all electronic components produce noise – but some produce more noise than others. The amount of noise that they produce depends not only on the type, construction and material used in the component, but also on the electrical conditions under which they are operated (ie, current and voltage) as well as, very significantly, the temperature. Noise is a particular problem in high-gain amplifiers where noise generated in the first stage receives the full benefit of the gain provided by all of the subsequent stages. We will be looking at this in greater depth next month.

Hum

Another problem that sometimes rears its head is hum. Hum is simply the appearance of a signal at mains supply frequency (or twice the mains supply frequency). Hum can be carried on supply voltage rails where it appears as a small AC signal superimposed on the DC supply. It can also find its way into an amplifier when stray magnetic fields (such as those that surround power transformers) induce current into nearby wiring. Hum can be reduced, if not completely eliminated, by good power supply design, using screened signal cables, and by adequate screening and grounding of chassis and chassis mounted components.

Sinewave testing

Fortunately, a quick inspection of an output waveform will usually provide you with a clue as to what the type of distortion (if any) has been introduced by an amplifier. First, of course, it is necessary to ensure that the input signal is free from distortion and this involves checking that it is a reasonably pure sinewave. Most signal generators are able to produce sinewave outputs with levels of distortion that cannot be discerned by the human eye when using an oscilloscope to view a waveform. As the level of distortion increases – typically to about 5% or more – the distortion starts to become visible in the form of a departure from a pure sinewave shape. This makes it possible to carry out a quick check on distortion by simply viewing the shape of an output waveform in response to an applied sinewave.

Sinewave tests are normally carried out at a frequency of 1kHz but, in order to verify low- and high-frequency performance respectively, these mid-audio band tests might be supplemented with performance checks at 100Hz and 10kHz. That said, there is rarely much need to carry out tests at any other frequency than that which coincides with the middle of an amplifier's frequency range and so 1kHz is invariably sufficient to carry out a quick distortion assessment.

Fig.4.5 shows various conditions. The undistorted sinusoidal input signal is shown in Fig.4.5(a). This is also the ideal shape for the output waveform which, if the amplifier is not introducing any distortion, should be perfectly sinusoidal. The effect of clipping is shown in Figs. 4.5(b) to 4.5(d). In the case of Fig.4.5(b) the positive edge

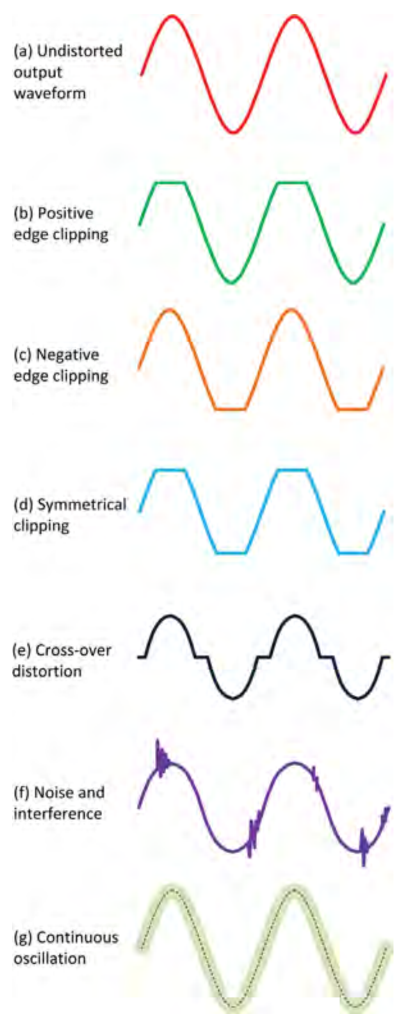


Fig.4.5 Using a sinewave signal to test for various distortion conditions

of the waveform has been clipped. Notice the flattening effect this has on the positive excursions of the signal. Fig.4.5(c) shows a similar effect applied to the negative edge of the waveform. These two conditions usually point to an incorrect bias adjustment where an applied signal becomes increasingly distorted (and clipped) as whenever the amplitude of the input signal exceeds a certain value. Symmetrical clipping (clipping of both positive and negative peaks) is illustrated in Fig.4.5(d). This condition usually results from applying an input signal of excessive amplitude. Reducing the amplitude below a critical value (below the point at which clipping starts to occur) will often correct the problem and reduce the distortion to an acceptable amount.

In Part 3 of the *Teach-In 2015* series we mentioned that cross-over distortion can be a problem when insufficient standing current is available in a complementary push-pull output stage (see *page* 42 of April 2015 *EPE*). Fig.4.5(e) shows this problem. Note how the output signal remains at zero until the input signal reaches a certain value, and how this form of distortion affects both positive and negative-going half cycles of the input waveform.

Apart from the repetitive waveform defects that we've seen thus far, a signal might also become contaminated by random fluctuations in signal level (noise) and sudden spikes of interference caused by switching and other transients induced into wiring (both internal and external). This problem is illustrated in Fig.4.5(f).

Finally, you might occasionally come across a circuit that, while intended to act as an amplifier, also acts as an oscillator! The reason for this is that, at some frequency (usually very much higher than the highest designed signal frequency) the internal phase shift becomes such that the design's intended negative feedback becomes positive, de-stabilising the amplifier and resulting in continuous oscillation. Designers of high-gain amplifiers usually try to avoid this problem by using only local feedback (ie, feedback over a single stage). In fact, we've used this technique in several of the amplifier designs that you've already seen. For example, C2 in our headphone amplifier (see Fig.3.8 on page 41 of April 2015 *EPE*) provides shunt feedback from the output of the first stage back to its input. In high-gain wideband amplifiers designed to operate at very high frequencies (eg, in the VHF/UHF radio spectrum) it can often be difficult to avoid parasitic feedback resulting from small amounts of reactance present within active devices, circuit wiring and the wire connecting leads of transistors and other components.

The typical effect of continuous (parasitic) oscillation is shown in Fig.4.5(g). Note that the amplitude of the parasitic oscillation often appears significantly reduced due to the limitations of the frequency response of the oscilloscope used to observe the distorted waveform.

Measuring distortion

The accurate measurement of distortion usually involves the use of some relatively sophisticated test equipment in the form of a spectrum analyser or wave analyser. When carrying out laboratory checks on our finished *Teach-In 2015* prototypes we used a wave analyser as well as a distortion factor meter. The former instrument is capable of measuring the level of the individual harmonics present in an output signal, while the latter provides us with a figure for the circuit's THD performance in a particular bandwidth (we usually restrict our measurements to an upper limit of 100kHz).

The distortion factor meter comprises a wide band AC voltmeter combined with a variable frequency notch filter that can be tuned so that it eliminates the fundamental frequency component present in the amplifier's output. In use, a reference level is set at 100%, with the filter switched out and then the level is measured again with the filter selected. In this condition, the signal measured by the voltmeter indicates the residual distortion present and its level (in relation to the fundamental) is displayed in decibels (dB) or as a percentage.

It is important to be aware that, when using a distortion factor meter rather than a more complex wave analyser, the instrument will respond to all in-band signals, including noise, hum and other non-harmonically related components. Nevertheless, this type of instrument can be very effective when carrying out a quick assessment of the performance of an amplifier.

Signal sources for accurate THD measurement

When working at very low levels of THD (typically less than 0.1%) the ultimate accuracy of the measurement

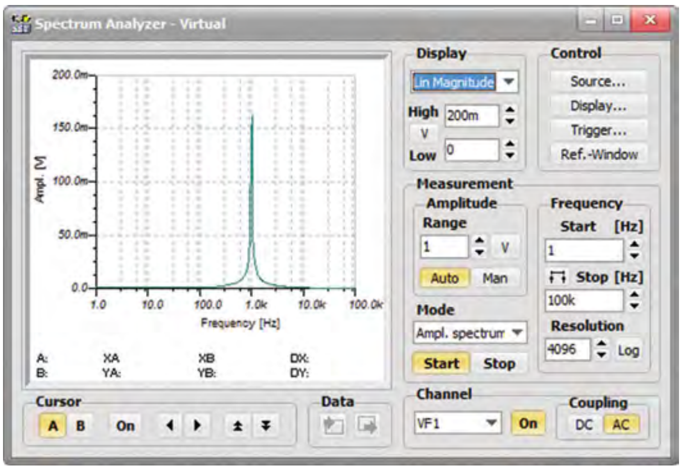


Fig.4.6 Output frequency spectrum analysis of the simple pre-amplifier described in Part 1

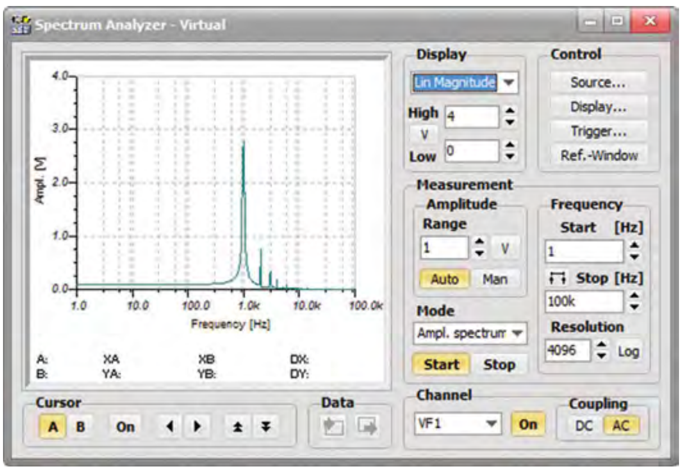


Fig.4.7 Appearance of harmonic distortion when the pre-amplifier is over-driven

becomes dependent on the quality of the input signal (which must be as near perfect as possible). We use a Radford low-distortion oscillator for our measurements. This instrument is capable of producing sinewave test signals with less than 0.003% THD. Note that popular low-cost function generators can often produce as much as 0.5% THD and so this type of signal source, although handy for the sinewave testing that we described earlier, is unsuitable for carrying out meaningful THD measurements.

Using virtual test equipment for distortion measurement

Computer-based virtual test equipment operating in conjunction with sound cards or a dedicated external analogue-to-digital converter (ADC) can also be used for distortion measurement. Measurements with virtual test gear will usually provide a reasonable guide as to likely THD performance, but care needs to be taken when adjusting signal levels and setting instrument parameters. For example, Fig.4.6 shows TINA Design Suite's Virtual Signal Analyzer being used to check the frequency spectrum of the single-stage preamplifier that we described in Part 1 of *Teach-In 2015* (February 2015 *EPE*). Notice how the 1kHz output signal has an amplitude of 150mV and that no other unwanted frequencies can be seen in the output spectrum. Increasing the input signal to produce an output amplitude of a little more than 2.5V results in the spectrum shown in Fig.4.7. Notice the appearance of second, third, fourth, fifth and sixth harmonics resulting from over-driving the pre-amplifier.

TINA's Virtual Signal Analyzer allows us to determine the amplitude of each of the individual harmonic components present in the output spectrum. For example, in Fig.4.7 the second harmonic has an amplitude of a little less than 0.8V and that of the third harmonic is approximately 0.4V. This information can be very useful at the design stage since it is possible to experiment with component values and input signal levels to check distortion level prior to prototyping the real circuit.

Distortion measurements can also be carried out later when prototypes of real circuits are available. Fig.4.8 shows a display produced by Virtins Multi-instrument when being used to check the sound card output and to ensure that the THD is sufficiently low to enable accurate low-level measurements to be carried out. As you can see, a great deal of information is available from the software. The virtual oscilloscope and signal analyser windows are simultaneously displayed (corresponding to the time domain and frequency domain graphs that we showed earlier in Fig.4.2). Notice that the frequency scale is logarithmic and that it extends from 20Hz to 20kHz so that it covers the full audio frequency range. Careful examination of the frequency spectrum shows the fundamental with an amplitude of 1V, together with odd-order harmonic components with descending amplitude. This suggest that the distortion arises from symmetric rather than asymmetric non-linearity (see earlier).

Various performance figures are reported in windows on the right of Fig.4.8. The first of these is the THD figure (0.0009%). This is extremely low. The next figure shows THD plus noise and hum (THD+N) and this would be the same figure that a distortion factor meter would report. The THD+N figure is, of course, greater than the THD figure, but it is still very acceptable at a mere 0.0015%. The third window shows the signal-to-noise ratio (SNR) and the reported value is a very acceptable 98.66dB.

The next window shows SINAD or 'signal in noise advantage'. SINAD is a measure of the signal quality that takes into account the presence of distortion. SINAD is calculated from the ratio of total output signal power (ie, signal power plus noise power plus the power arising from distortion components) to the noise-plus-distortion power. As might be expected, the SINAD performance is only slightly worse than the SNR performance, but the SINAD figure is usually a more reliable measure of performance, particularly when a significant amount of noise and distortion is present in a system.

The last window shows the effective number of bits (ENOB) which provides an indication of the dynamic performance of the system in relation to that of an analogue-to-digital converter (ADC). Since the resolution of an ADC is specified by the number of bits used to represent an analogue quantity, the number displayed indicates the resolution of an ADC that would operate with the same resolution as the circuit under consideration. The figure quoted here is just less than 16-bits and is indicative of a high-quality audio system.

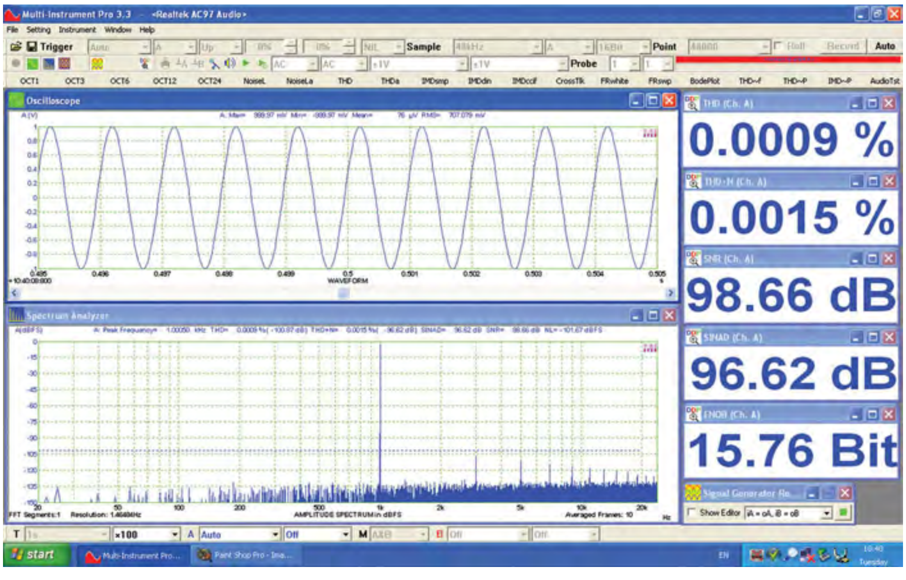


Fig.4.8 The Virtins Multi-instrument display provides a variety of useful distortion measurements

Discover: square wave testing

As mentioned in the main text, in addition to its use for a rapid check on distortion (or the lack of it), an oscilloscope can also be used to provide a rapid assessment of the frequency response of an amplifier or other electronic system. Instead of using a sinewave as an input signal, a square wave input is used. Square waves comprise an infinite number of sinusoidal harmonic components added to the fundamental sinewave and so any defects in the frequency response of an amplifier will show up very quickly from an examination of the shape of the waveform of the output signal when a square-wave input is applied. As a result, it is possible to assess whether the frequency response is good or poor (a perfect square-wave output would correspond to a perfect frequency response).

Square-wave testing can provide a rapid assessment of the frequency response of an amplifier. A perfect square-wave output, Fig.4.9(a), indicates that the amplifier under test has a flat frequency response. Figs. 4.9(b) and 4.9(c) are typical of an amplifier having a good high frequency response coupled with a poor low frequency response, while Figs. 4.9(d) and 4.9(e) are indicative of poor high-frequency response and good low-frequency response. The damped oscillation, or *ringing*, shown in Fig.4.9(f) occurs when the amplifier's step response (its response to a sudden and very rapid change in signal level) produces momentary oscillation. This can occur where significant inductive reactance is present at the same time as capacitive reactance. The combined effect of these two 'opposite' reactances results in resonance, where the sudden changes imposed by the rising or falling edges of the square wave cause energy to oscillate backwards and forwards between the two opposite reactive components. The transfer of energy between the two components decays (due to resistive

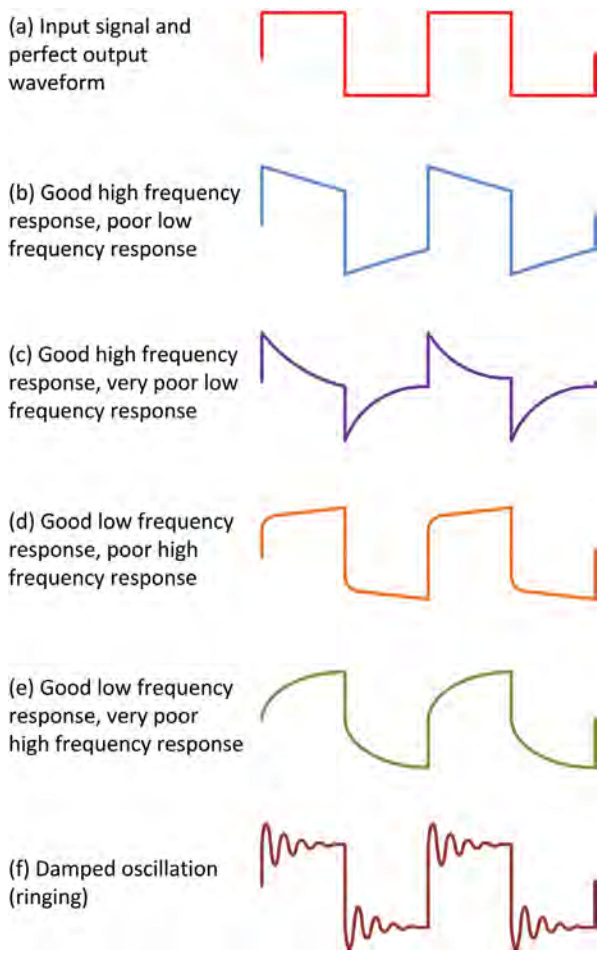


Fig.4.9 Square wave testing can provide a rapid means of checking the frequency response and transient performance of an amplifier

losses) and so the oscillation quickly (relative to the period of the applied square wave) settles to a steady value at one or other extreme of amplitude.

Get Real: testing the simple headphone amplifier

Our second *Get Real* project, a simple headphone amplifier, was designed to satisfy the need for a simple means of boosting the output of an MP3 player, mobile phone, tablet, or similar portable device when used with high-quality headphones or a small loudspeaker. Our circuit design used just three transistors; a single-stage common-emitter driver and a complementary output stage.

The headphone amplifier was simulated during the design phase using Circuit Wizard to check the circuit, verify the DC conditions, and to produce the PCB layout for our prototype amplifier. We then used TINA Design Suite to carry out accurate measurements of the circuit performance (voltage gain, input and output impedance, output power) and to select component values for C1 and C2 that would provide the desired frequency response (10Hz to 20kHz). Our simulated circuit is shown in Fig.4.10. Note that we have added a 600Ω load to the output.

As with our other *Get Real* projects, we built the prototype circuit and tested it using both virtual and real test instruments in order to verify the design and to see how close the performance of the prototype circuit came to the original design specification. The results of our measurements are shown in Table 4.1. Our test measurements were carried out using a 600Ω source impedance and a load that was variable from 8Ω to 600Ω.

For stereo operation, separate amplifiers will be needed for the left and right channels. The left and right output signals should be taken from CN2 to a suitable off-board jack connector (in which case the board-mounted 3.5mm jack connectors on each board are not required). For mono operation, the left and right channels need to be fed to the input via equal value resistors of typically between 1kΩ and 10kΩ. This arrangement is all that is required to sum the two signals and produce a corresponding mono signal at CN1.

Waveform and distortion testing

The simple headphone amplifier makes an excellent platform for checking out the sine and square wave-based distortion and frequency response measurements that we described earlier. Simply select the appropriate input waveform, a test frequency of 1kHz, and ensure that the signal amplitude is not so great that it over-drives the amplifier (we recommend an input signal of no more than about 100mV). By varying the input signal level, output load impedance, and the two pre-set adjustments, you will be able to investigate cross-over distortion, symmetric and asymmetric clipping under both simulated and real-world conditions; here are some ideas:

1. Adjust the amplifier as described on page 42 last month and connect a 600Ω load resistor to the output. Check that VR1 and VR2 are correctly set for a symmetric and undistorted output waveform. No clipping or cross-over distortion should be discernible.
2. Set the input amplitude to 100mV using a sinewave signal at 1kHz. Set VR2 to minimum resistance and, if necessary, adjust VR1 for a symmetric output waveform. The cross-over distortion should be very noticeable.
3. Next, increase the signal amplitude to 200mV and adjust VR1 for a symmetric output waveform. The cross-over distortion will still be present but will be less noticeable.
4. Increase the input signal to 400mV and set VR1 to minimum. The output waveform should be noticeably distorted with negative edges clipped.
5. Keep the input signal at 400mV and set VR1 to maximum. The output waveform should remain distorted, but now the positive edges should be clipped.
6. Finally, increase the input to 800mV and adjust VR1 for a symmetric output waveform. This should show symmetric clipping (note that the amplifier becomes over-driven when the input voltage exceeds 400mV).
7. Use the Virtual Spectrum Analyzer to investigate the frequency spectrum of the clipped waveform. We recommend that you use the TINA settings shown in Fig.4.12.

Table 4.1 Measured performance of the simple headphone amplifier

Specification	Design objective	Simulated (TINA)	Measured
Mid-band voltage gain	10	7.3	10
Input impedance (1kHz)	5kΩ	7kΩ	5.5kΩ
Frequency response	10Hz to 20kHz at –3dB into 600Ω	2Hz to 25kHz at –3dB into 600Ω (see Fig.4.11)	10Hz to 20kHz at –3dB into 600Ω
Output power for 1% THD	50mW into 15Ω at 1kHz	Not measured	20mW into 15Ω at 1kHz
Distortion	<0.2% THD at 1mW into 600Ω at 1kHz	Not measured	0.3% THD at 1mW into 600Ω at 1kHz
Supply current	<10mA with no signal (9V DC supply); <50mA at max. output	4mA with no signal	5mA with no signal; 34mA with 50mW output into 15Ω at 1kHz

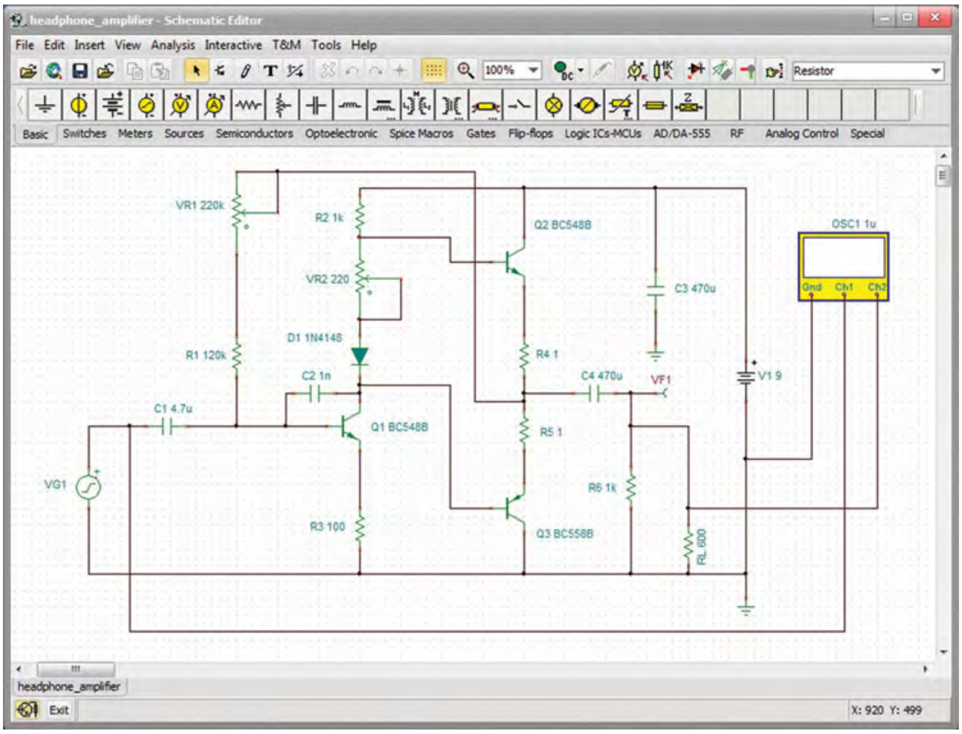


Fig.4.10 The headphone amplifier simulated using TINA Design Suite

(Fig.3.12 in page 42 April 2015 *EPE*). The superimposed layer results in the five track breaks shown in Fig.4.13(a). The correct artwork is shown in Fig.4.14 but if you have already manufactured the board by photocopying the layout you will only need to add the five small solder bridges shown in Fig.4.13(b). Apologies for any problems this has caused. Note that the board supplied by the *EPE PCB Service* is correct and does not need any modification.

Next month

In next month's *Teach-In 2015*, *Knowledge Base* introduces the use of filters to modify the frequency response of an amplifier. *Discover* takes a detailed look at noise and the means by which it can be reduced. Our practical feature, *Get Real*, describes the design and construction of a versatile tone control stage that can be used on its own or in conjunction with the other projects featured in our *Teach-In 2015* series.

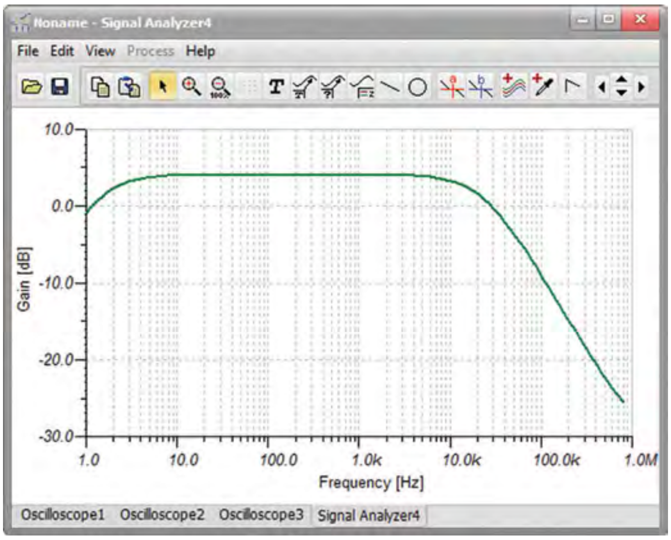


Fig.4.11 The frequency response simulated using TINA

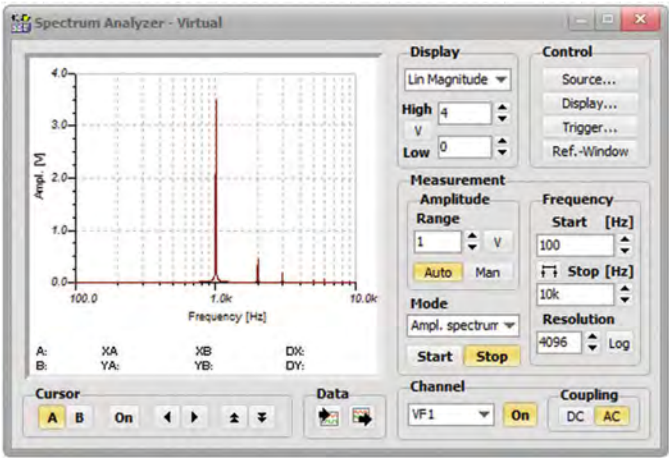


Fig.4.12 Recommended TINA Design Suite settings for carrying out the spectrum analysis

If time permits, it is worth repeating these measurements with load resistances of 50Ω and 15Ω. A significant increase in distortion should be noticed due to the corresponding increase in output power.

Errata

Unfortunately, the silkscreen outline for the (optional) jack connector used in our simple headphone amplifier has also appeared on the PCB track layout

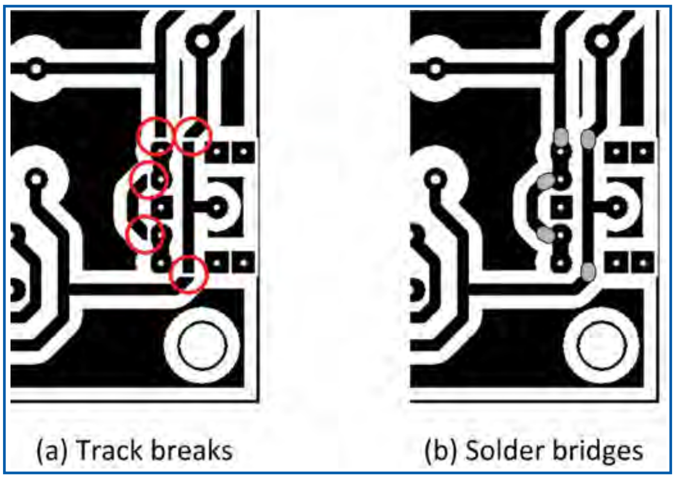


Fig.4.13 Track breaks and solder bridges

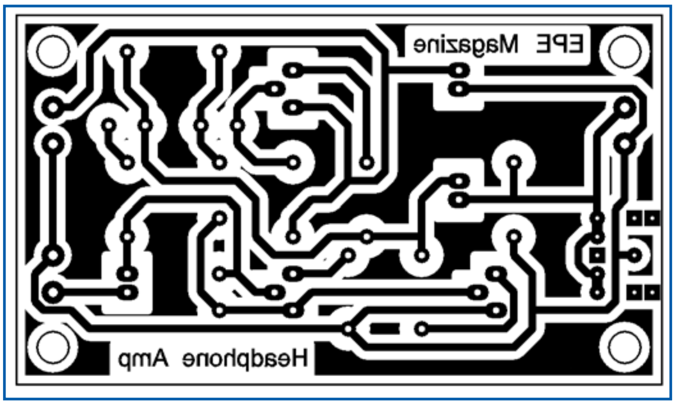


Fig.4.14 Corrected PCB track layout

NET WORK

by Alan Winstanley

Getting more fibre

SOME fifteen years have elapsed since I sang the praises of ADSL, the high-speed broadband service that promised vast benefits over traditional 56k dial-up Internet access. At the time, BT heralded the launch of ADSL by urging web designers and developers to provide content optimised for this new high-speed Internet access. At a business presentation, BT speakers craved for video and multimedia websites, Flash presentations, music and more – anything that could harness the forthcoming ADSL service effectively. The higher bandwidth was suddenly there, but the killer applications were not.

This was against a background of BT throttling the roll-out of ADSL on its network to begin with. Needless to say, cities and urban areas were the first to benefit from higher speed connectivity, while the provinces were subject to a lottery system of 'trigger levels' – the point of interest after which BT might or might not upgrade a local exchange to offer 512kbps ADSL. Eventually the rollout accelerated and the speeds and coverage increased, with advertised bandwidth ratings always qualified with the weasel words 'up to...' in front of them.

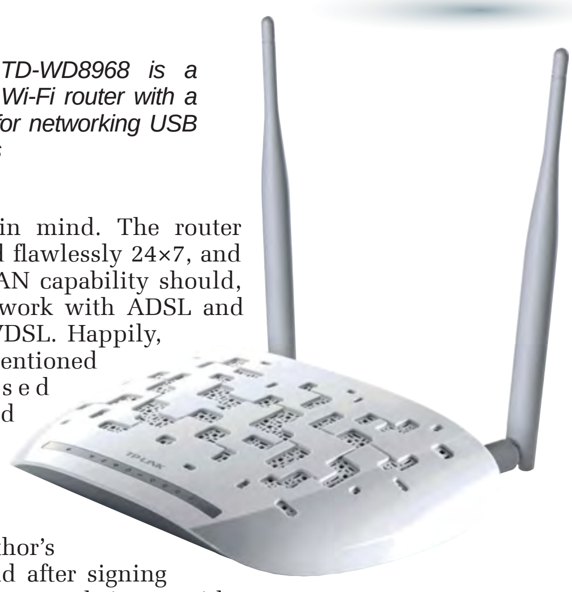
With ADSL now covering much of the country, fast forward to the present day and ADSL is proving too slow and stuttery for many users. Countless mobile phones, tablets, Smart TVs and other devices are hooked to a network; and movies on demand, catch-up TV programs and streaming Internet services now gobble bandwidth galore.

The next stage in the Internet's evolution involved the expansion of fibre-optic-based services. Using 'Fibre to the Cabinet' (FTTC), download speeds of 50-80Mbps or so are now attainable, depending on geography. However, for engineering reasons some rural users experience speeds that are not much better than ADSL to begin with, always assuming they can purchase FTTC anyway. Hence, groups such as B4RN (Broadband for the Rural North) have taken matters into their own hands by funding the local delivery of fibre-based networks across rural areas themselves. Other options for wireless mesh networks or satellite may become available to Britain's more remote outposts.

FTTC still relies on ancient copper wires, junction boxes and rickety overhead cables to deliver superfast broadband the last few hundred metres to the home, but in today's economy it is the most practical and realistic way of incrementally squeezing more capacity out of a creaking system. Fibre can prove cheaper than ADSL, and for now it's all there is, but as a matter of course the next gigabit generation will need a new infrastructure based on FTTH/FTTP (Fibre to the Home/Property) to deliver 4K Ultra HD streaming video to our homes. Perhaps that is something *Net Work* can cover in the next fifteen years!

A superfast switch

Fed up with throwing away cheap consumer ADSL routers after two years, my sturdy well-ventilated Billion 7800N router, mentioned in *Net Work* April 2011, was bought for my busy office with one eye on a future upgrade to fibre hopefully in mind. The router has worked flawlessly 24×7, and its dual-WAN capability should, in theory, work with ADSL and superfast VDSL. Happily, the aforementioned fibre-based broadband finally arrived last week in the author's locality, and after signing up for a free upgrade it was with great anticipation that I plugged my Billion 7800N router into a shiny new VDSL modem. The moment of truth arrived – would my four-year-old dual-WAN router deliver on its promises? Simply by logging in and pointing the router to EWAN instead of ADSL, the router then rebooted and my system awoke to experience the new era of superfast broadband without so much as a yawn. The joys of VDSL were soon realised, with snap-action surfing and a torrent of digital data darting over my system like never before.



TP-Link's TD-WD8968 is a dual-WAN Wi-Fi router with a USB port for networking USB peripherals

Even so, one local user claimed that she did not notice much difference at night-time, though a speed check at **www.broadband.co.uk/broadband-speed-test** showed variations from 50Mbps down to 7Mbps periodically. The system optimises itself over several weeks and speeds are holding up thus far; a network will only go as fast as the slowest bottleneck, and it will be interesting to see how well the service holds up at peak periods. If readers at home or abroad have any practical experiences to share about upgrading to superfast broadband, please feel free to email me at the usual address and we'll consider them for *Readout*.

Readers who are in the market for a new router might consider a future-proof dual-WAN type for flexibility, but bear in mind that a VDSL modem is needed to connect directly to the incoming master socket. Some routers now have a VDSL modem built-in, a point worth watching for, otherwise a separate VDSL modem with mains adaptor is also needed. BT Openreach can provide them.

Routers have many features and are very flexible these days, which makes comparing and choosing a new one as complicated as ever. The Billion dual-WAN router offered four gigabit ports, plus a fifth port which doubles up either as an EWAN connection or a handy 100Mbps port instead. This dual-purpose option appears on other routers, such as a TP-Link TD-WD8968 that I recently installed in a small office. This nifty Wi-Fi router also has a USB port

The rear of the TD-WD8968 showing the ADSL, USB and power connections. The LAN4 port doubles as an Ethernet or WAN (fibre/cable) port

for hosting, which could be used for printer sharing on a network, or a 3G USB dongle, maybe as a 'mobile phone' style data backup in case the landline or broadband go down. The WD8968 will also act as an FTP or media server and is remarkable value at £29. When upgrading, do check that a router is IPv6 compatible, and ideally choose one with a built-in VDSL modem for direct connection to fibre.

Having lost a port to the EWAN connection, my own router was now 'maxed out', so the next question was how to add more Ethernet ports to my network. The answer was to add a cheap-and-cheerful unmanaged switch. These small add-ons are as easy to use as a powered USB hub and simply require a short Cat5 or Cat6 Ethernet lead connected to the router (thereby taking up another precious port!). A basic 5-Port switch such as the compact steel-clad TP-Link TL-SG105 will provide a further four gigabit Ethernet ports allowing, in my case, a total of seven devices to be added. Eight-port or more are available if higher capacity is needed. Many switches have keyhole slots to hang them under a desk out of the way, but of course they all need yet another mains adaptor.

Switches can cost as much as a cheap router, with little to choose between different brands, but preferably choose a gigabit version over a 10/100 type, which will be listed for as little as £10. Finally, a selection of Ethernet cables is always available on eBay and I find it's worth keeping a few different lengths of 'snagless' cables in stock. Some different colours brighten the day!

Not so tasty spam

Unsolicited mail continues to be the bane of every Internet worker's life. In a recent case that came my way, office workers suddenly started to receive strange 'reject' emails that resulted from an email campaign sent with a forged (or 'spoofed') email address as the sender. Spam email was bouncing back to the genuine owner of the email address who was suddenly deluged by a lot of rejected mail (called 'back scatter'). The spam was being sent out through their ISP's server and was authenticated to appear genuine, in an effort to avoid some spam filtering systems. One contributing factor was the possibility that the firm's email account password was the same one used to log into various websites: if a remote web server had been hacked then this account login information might have been stolen and sold to spammers. Armed with this information the spammers could then impersonate the firm and broadcast unsolicited mail via their ISP's mail servers. (The spammer was traced to Taiwan.) Alternatively, a Trojan or virus was residing on the office computers, sending out spam blindly without the company's knowledge.

The first job is to immediately change the email passwords, and then scan each PC on the network using Malwarebytes

TP-Link's TL-SG105 is a typical compact unmanaged 5-port gigabit switch that enable more ports to be easily added to a router. This one is compact, steel clad and mains powered. LED colours indicate the speed.

Anti-Malware from **https://www.malwarebytes.org**. An anti-virus scan will also be wise. After that, it is simply a case of sitting it out and waiting for the tidal wave of spam to subside. Of course, this causes nothing but inconvenience and the 'opportunity cost' is also high: office staff could be doing something more profitable, backlogs build up and important emails that are buried in a deluge of spam could be overlooked or deleted accidentally.

Over the past 20 years, many traditional Internet users have grown up with the principle of a POP3 or IMAP mailbox, but plenty have given up on the idea due to the headache of spam and viruses, and they have migrated their email to web-based services like Outlook, Gmail or GMX instead. A number of choices were suggested last month. A major benefit is that web-based mail can be handled by mobiles, tablets or notebooks, avoiding the need to run an email client program tied to a PC.

The thorny topic of spam also surfaced recently when I noticed my own email service suddenly dried up. I suspected my external anti-spam filtering service was having problems: it supposedly fetches all my POP3 mail, screens out any spam and leaves the remainder for me to collect as normal. Some test emails sent to myself proved that mail was disappearing out of my POP3 box, but it was not being filtered ready for collection and my email was therefore piling up somewhere else. If available, an ISP's webmail service can provide a helicopter view of what's going on at times like these, or a service such as **http://mail2web.com** allows users to track the progress of mails via a web browser. Unfortunately, the spam-filtering firm fell silent and evidence soon mounted that some email had been lost altogether. A batch of mail suddenly surfaced out of the blue, a fortnight later – better late than never! – but I decided to disable the spam-filtering service altogether and handle the problem of filtering spam myself.

Washing your mail

Over the years, only a handful of companies offered to filter an individual POP3 mailbox this way – fetching a mailbox, filtering it and leaving the rest to await collection. This led to another problem, as a lot of googling revealed that almost every other anti-spam service required changes to the domain's MX (Mail Exchange) records, so the whole domain's email feed was directed to them at root DNS level. This puts your email entirely in their hands. They were mostly enterprise-level blue chip solutions for business and a typical professional spam filtering service such as Mail Essentials from GFI.com costs £216 a year. The 'gotcha' is that a minimum of ten users must be ordered at £18 + VAT each.

With paid-for anti-spam services being prohibitively expensive for a single user, for the time being I decided to try the popular **Mailwasher** software, which can be downloaded free from New Zealand's Firetrust Ltd at: **www.mailwasher.net**.

This extremely handy free Windows software (Linux and Mac versions are promised) fetches email from an individual POP3 mailbox and then sets to work screening it. Learning as it goes along, some mail will automatically be recognised as spam and flagged, while other mail can be given the thumbs-up or thumbs-down by the user. Addresses can also be blacklisted or whitelisted. The net result is that a busy mailbox can quickly be scanned and pre-filtered, and by clicking the 'Wash Mail' button the junk can then be removed from your mailbox. With your POP3 mail given a rinse, your regular email program then fetches the remainder. Mailwasher can also bounce dodgy mails back from whence they came (or appeared to, anyway) – a feature best used with care and consideration.

The free version will filter a single POP3 mailbox, while Mailwasher Pro handles multiple mailboxes, syncing across mobile devices and has a larger preview window and Recycle bin. A one-year Pro license for three PCs costs £24.95, or a lifetime license is £64.95 – not exactly a trivial cost, but perhaps worthwhile if you are inundated with junk mail.

Busy Microsoft Outlook users might also like Firetrust's OkInBox plugin for Outlook. It claims to allow users to prioritise important emails and places everything else into a

Mailwasher is free software that pre-filters an individual POP3 mailbox. Note the Friends and Blacklist address books on the right

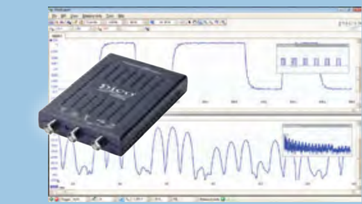
digest, which is sent to you as a newsletter to be dealt with at a more convenient time. The author has not tested it.

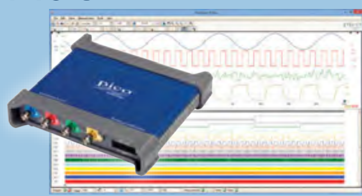
Lenovo Superfish

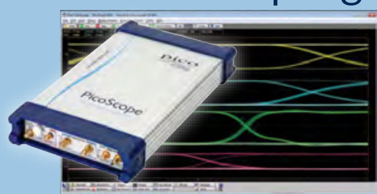
The PC manufacturer Lenovo (strapline 'For those who do') hit the headlines in February when it was revealed that Lenovo was indeed 'one that had': a number of its notebooks were pre-installed with a sneaky adware system known as Superfish. According to Lenovo, the Chinese brand that acquired IBM's desktop PC and laptop business, Superfish 'assisted customers with discovering products similar to what they are viewing.'

This sounds innocuous enough until the underlying *modus operandi* of Superfish is revealed. Superfish is an image-based product placement and promotion system that perpetrates what is often called a 'Man in the Middle' attack. It intercepts secure https:// web traffic and injects it with its own advertising images. While users may think they are reaching out to a secure website, in fact they are not: the Superfish adware breaks the end-to-end tunnel and acts like a man in the middle without the user's knowledge. Users are unconsciously interacting with Superfish residing on their system, not with the remote website, and it plays tricks with secure certificates by certifying itself as a trusted Certificate Authority (CA). Thus it fools the system into thinking that communications are secure and trusted when they are technically not. Lenovo disabled the server-side interactions to cripple it at the start of the year and belatedly offered a Superfish removal tool, but its reputation has suffered great embarrassment. If you have a Lenovo notebook then visit: **http://support.lenovo.com/en/product_security/superfish** to check whether yours needs checking over.

That's all for this month's *Net Work*. You can email the author at **alan@epemag.demon.co.uk**







Obtaining and storing components

I think it is fair to say that obtaining all the components for an electronic project has never been entirely straightforward. A wide variety of parts are used in most projects, ranging from mechanical items to highly specialised and hi-tech semiconductors. It would be nice and convenient to purchase all the components from a single supplier, but in most cases it is necessary to use two or more sources.

Buying the required bits and pieces has probably become more difficult over the years, with the range of available components becoming ever wider and more diverse. The number of suppliers catering primarily for amateur electronics enthusiasts has diminished, which has exacerbated the situation. Fortunately, the very large suppliers of electronic components to commercial and educational users, such as RS and Farnell are happy to accept credit/debit card orders from customers who do not have an account with them.

Netting components

The Internet is an important tool when obtaining components, and getting everything you need could be very difficult or even impossible unless you have Internet access. I would guess that every electronic component supplier has some form of online catalogue and ordering system, plus search facilities to enable the required parts to be located very quickly. A normal search engine can be useful when trying to locate sources of supply for the more elusive components. I find that eBay is an increasingly useful supply route, and the range of electronic parts on offer there is enormous, especially for deleted or obsolete parts.

It is worth repeating the much-given warning that it is best not to buy any components for a project until you are sure that there are reliable sources for *all* the components. This is especially important when building a project that was published over a year ago. With some components it is almost literally a matter of 'here today – gone tomorrow', because the remaining stocks of discontinued components often sell very quickly.

Mundane components such as resistors, capacitors and common semiconductors should not be a problem, and it is the more specialised parts that can be problematic. Many projects are based on an integrated circuit that is dedicated to one specialist function, or perhaps have an unusual sensor of some kind. It is components such as these that are most likely to give supply problems. Again, the Internet can be useful for finding elusive components, but only if they are still actually in stock somewhere! Components that were discontinued long ago do occasionally make an appearance on Internet auction sites, but there is no guarantee of success even if you are prepared to wait it out.

Stocking up

Ordering every component for a project, right down to the last nut and bolt, can be quite time consuming. It obviously depends on the size of the project, and in general the situation is better than it was in the past. Modern electronic circuits tend to be based on a microcontroller or a specialist integrated circuit (IC), with the number of discrete components being kept to a minimum. The average component count is much lower than when things were done the hard way using circuit boards that contained long rows of resistors and capacitors.

Even so, ordering a full set of parts can be a time-consuming process. There can be delays if one or two of the components are out of stock, or if you overlook something when ordering the components and have to place a further order. Before too long, many project builders opt to have a stock of the more frequently used components, so that it is no longer necessary to buy every single component each time a project is constructed. All or most of the more mundane components can then be taken from the stock of parts, with the expensive and more specialist items still being purchased in the normal way. Fortunately, the most-often used components are generally the cheapest, such as resistors and capacitors, so buying a good stock of parts does not have to be expensive.

Low-wattage resistors are used in significant numbers in practically every electronic circuit, and are the cheapest of electronic components. This makes them a good starting point when putting together a stock of components. The more exotic resistors, such as close tolerance (one per cent) or high power types are more expensive and used relatively little. It is probably best to buy them as and when necessary. Most projects only need 'cheap as chips' carbon film resistors with a power rating of about 0.25 to 0.5W and a five per cent tolerance rating. These are the type that should be obtained for stock. These days, metal film resistors, which are higher in quality than the carbon variety, seem to cost little extra. They are a better choice if they can be obtained at a price that suits you.

An obvious problem when buying a stock of resistors is the large number of values in use. In the E12 series of values there are some 85 values from 1Ω to 10MΩ, and more or less double this figure in the E24 series. Table 1 shows the available values in each series. These are the basic values, but resistors are available in one tenth and one hundredth of these figures, and ten times them, one hundred times them, and so on. They are known as 'preferred values'. Although at first sight the preferred values might seem to be a slightly odd selection, they are quite logical. The E12 and E24 series respectively go up in increments of about 20 and 10 per cent, albeit with some compromises because only two digits are available per basic value. For close-tolerance resistors there are actually E48, E96 and E192 series, although these seem to be little used in practice. Preferred values are also used for capacitors, but with some types of capacitor, and the electrolytic type in particular, it might only be values in the E6 range (see Table 1) that are available.

Most projects only require resistors with values from the E12 series. Initially anyway, it is probably best to concentrate on the E12 range and not bother with the additional values in the E24 series. There is a slight complication caused by some values featuring in circuit after circuit, while others are used to a much lesser extent. In particular, values such as 1kΩ, 10kΩ and 100kΩ are used much more often than the ones immediately above and below these values. It makes sense to obtain more of the popular values, and fewer of the ones that are used less often.

Weighting game

The quick and easy way of obtaining a stock of resistors in a wide range of values is to buy one of the resistor kits or development packs that are available from some suppliers. These typically provide a full range of values from 10Ω to 1MΩ, or in some cases a more generous 1Ω to 10MΩ. The very high and very low values are little used in practice and for this reason are sometimes omitted from resistor kits. Apart from this, in most cases there is no weighting of the quantities to match the popularity of each value. However, you can provide your own weighting by obtaining extra supplies of the most popular values, such as 1kΩ, 4.7kΩ, 10kΩ, 47kΩ and 100kΩ.

Table 1

E6	E12	E24
10	10	10
-	-	11
-	12	12
-	-	13
15	15	15
-	-	16
-	18	18
-	-	20
22	22	22
-	-	24
-	27	27
-	-	30
33	33	33
-	-	36
-	39	39
-	-	43
47	47	47
-	-	51
-	56	56
-	-	62
68	68	68
-	-	75
-	82	82
-	-	91

The quantity of each value is often quite generous though, so there might be little point in buying any extra resistors. It is more likely that there will be some components that will never be required. While this is a bit wasteful, it is unlikely to represent a waste of money. With resistor kits you are bulk buying and the cost per resistor is usually very low. Even if half of them are never used, it is still likely to cost much less than buying resistors in small quantities as and when they are needed.

Development kits should not be confused with 'bargain assortments', 'bargain bags' and the like. With a development kit you know exactly what you are getting, which should be a reasonably comprehensive range of E12 values. The various bargain offerings tend to contain a range of values that has large gaps, and there may well be a lot of values from the E24, E48 and E96 ranges. Although they are generally well worth the money, the bargain offerings are better suited to experienced users who can sort out the good from the bad or indifferent.

Doubling up

An alternative way of building up a stock of resistors is to buy two or three times as many of each value as you actually need for each project. This approach has a couple of advantages, and one of these is that no initial outlay is required, and you will hardly notice the slightly higher cost when ordering each set of components. The second advantage is its built-in weighting. The buying of each value is based on the quantities used in projects of the type you build. While this method cannot forecast your future needs with absolute precision, it should be reasonably accurate in this respect.

Although this method works quite well, it has the obvious drawback that it could take a long time to build up a large stock of resistors. Also, as you are buying the components in fairly small quantities it would probably be relatively expensive in the long term. This method can be used with other components, such as inexpensive capacitors and semiconductors, but it is likely to be a waste of money if applied to any components that are expensive or specialised.

High capacity

It is very useful to have a stock of capacitors, but it is more difficult and expensive than producing a stockpile of resistors. As with resistors, the range of values in common use is very wide, and if anything it is even wider than for resistors. The cost of the more commonly used capacitors is not very great, but keeping the cost of a useful stockpile within reason could be difficult. Perhaps the biggest problem is that there are numerous types of capacitor for any

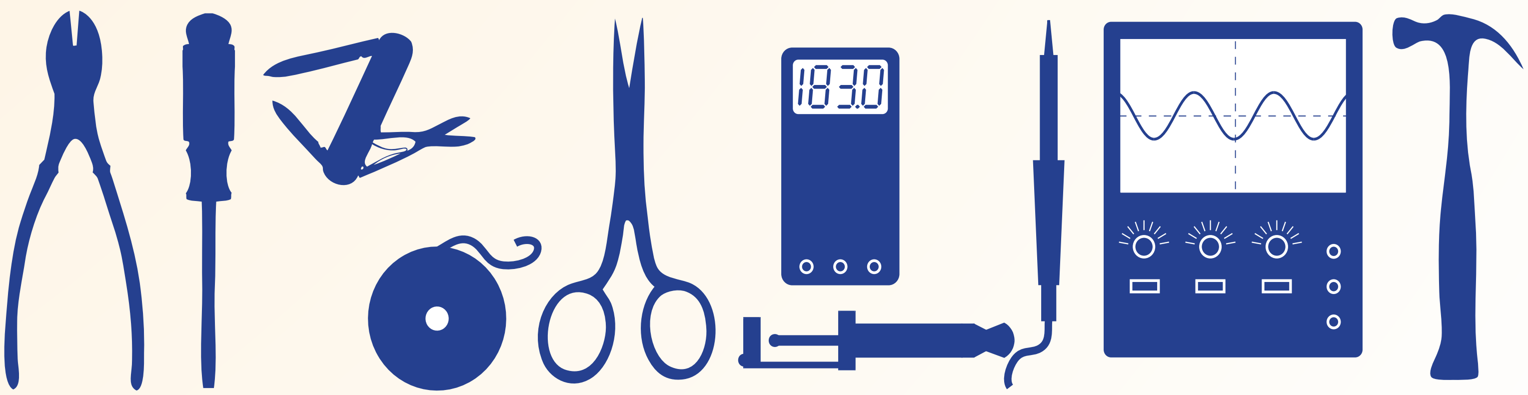


Fig.1. Capacitors are available in a wide range of values, types and physical sizes. These are electrolytic (top left), ceramic (top right) and plastic foil (bottom) types. It would be impractical to have a stock of all types and values, but it is worth having a reserve of the commonly used types

given value. There are fewer types at the higher and lower values, but in between there could easily be ten or twenty types to choose from. Fig.1 shows a selection of capacitors, but only a fraction of the available types are represented here.

Matters would be easier if the perfect capacitor was available at a low price. With the current state of capacitor technology, the choice tends to be between expensive components that perform well in a wide variety of situations, and low-cost types that have important limitations. Circuit designers have to select a type of capacitor that has suitable characteristics for each application, and is also cost effective. One application might require high accuracy and stability, while another might need effective operation at very high frequencies, and a third might need a component that can operate at high voltages. Many applications are not particularly demanding, and practically any capacitor of the right value will suffice. However, unless you know what you are doing it is advisable to only use a capacitor of the type specified in the components list.

A further complication is that most capacitors are now of the printed circuit mounting (PCM) variety, and these usually have very short leads. In fact, they have something more like the pins of an integrated circuit rather than anything that could really be termed a lead. PCM capacitors of a given type and value can, and often are, available with two or more pin separations. It is sometimes possible to manoeuvre a component into position on a circuit board even if it has the wrong pin spacing. However, doing so is unlikely to give neat or reliable results, and could result in the component being damaged.

Choice selection

Putting together an all-inclusive stock of capacitors is not really a practical proposition. It would have to contain a vast number of components and would be very expensive. On the other hand, it can be worthwhile putting together a well-chosen stock of the more common capacitors. Electrolytic capacitors are a good starting point because only a limited range of values are used to any extent. Some components in what might be termed the E3 series (1, 2.2, 4.7, 10 etc) from 1µF to 100µF should suffice. Modern circuits operate at low voltages, so any voltage rating from upwards of 25V will suffice. The lead spacing of electrolytic capacitors is less important than normal because these components do seem to have proper leadout wires. Provided you obtain modern miniature components they should fit quite easily into practically any component layout.

Low-value capacitors were used quite extensively in the days when building radios and other radio equipment was very popular, but they do not feature a great deal in modern electronics. Some polyester or polycarbonate capacitors in the range 1nF to 470nF are likely to be more useful. A few components of each value in the E12 range and with a couple of different lead pitches would be very useful, but would not be cheap. It is possible to obtain capacitor kits and development packs, but these mostly contain one type of capacitor in a relatively limited range of values, rather than a broad range of values and types. It would probably be necessary to obtain at least two or three different types in order to produce a really useful stock of capacitors. Development packs usually represent a very cheap way of buying capacitors, so for my money this almost certainly represents the best approach.

Bits and pieces

Resistors and capacitors are the best components to stockpile, but there are a few other components that are worth adding to your hoard. General-purpose silicon diodes such as the 1N4148 or 1N914 are well worth having, and it is probably worth including some small 1N4007 rectifiers. These can be used in place of any other rectifiers in the IN400* series. Miniature toggle switches are likely to be useful. A large reel of solder is probably the most useful stock item. Do not overlook common items of hardware such as nuts, bolts, spacers, and stand-offs. This type of thing tends to be sold in packs of ten or more, so you will probably end up with a stock of them whether you want it or not! If you find that you buy any component very frequently and it is not expensive, then it is probably worth having some in your stockpile of components.

Safe storage

It is important to store components properly, rather than just throwing everything into a couple of large crates. Having everything in one or two boxes makes it difficult to find the part you need, and results in wear and possible damage to all the components. Compartmentalising the components makes it much quicker and easier to find whatever you need. Making or improvising your own storage system is a fairly easy task, and there are plenty of ready-made miniature chests of drawers available at quite low prices. Having a separate drawer for each value is perhaps taking things too far, but a separate drawer or compartment within a drawer can be used for groups of similar values, such as 100Ω to 820Ω. It helps if things are arranged logically and each drawer has an appropriate label.

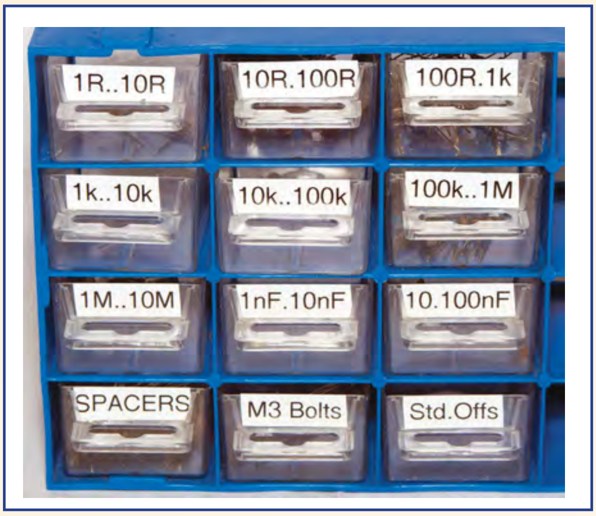


Fig.2. In order to make it quick and easy to use, a stock of components should be well organised. Suitably labelled miniature chests of drawers are ideal for storing components in a properly organised fashion

PIC n' Mix

Mike Hibbett

Our periodic column for PIC programming enlightenment

The Internet of Things

OUR articles on the development of a low-power microcontroller board over the last year have been heading in a particular direction – towards what (I hope) is one of this decade's technological revolutions – 'the Internet of Things.' The Internet of Things, now commonly referred to simply as 'IoT', is about the development of small, low-power sensors that can communicate via the Internet (or 'The Cloud', as it now seems to be called).

What is the Internet of Things?

Any electronic device that can communicate with other devices over the Internet could be classified as an IoT device, but the term is more typically used to refer to small sensor-equipped devices. These sensors can be anything, from the usual temperature and humidity sensors in the home, lamp switches or heating controls, right through to industrial control measurement and reporting systems and even vehicle tyre pressure monitors. The defining properties of such a device are small size, communication capabilities and generally, with a few exceptions, ultra-low-power consumption. And when we say 'ultra-low power', we are talking about micro-amps average current consumption, with peak requirements (in the brief time that a message is being transmitted) in the order of a few tens of milliamps. These requirements are driven by the capabilities of the power source; typically, IoT devices are expected to be powered by a tiny coin-cell battery that can supply sub-milliamp currents continuously and support a short-term peak load of a few milliamps. Almost every day it seems a new IC is released supporting this kind of power profile, and it's no surprise – in a near-future world where our homes are littered with tiny, self-contained computing devices, we wont want to be changing batteries every other week. We, the consumer, will want to fit a battery and then forget about the device for a year or two. And of course, we expect the device to communicate its battery life so we know weeks in advance when it is about to fail.

Companies working on IoT devices promise all of these things. Let's hope reality matches the hype!

Wearable computing

There is another technological revolution occurring at the moment, in the IoT field called 'Wearable computing'. It would be very surprising if you had not heard of Google Glass or smart watches, devices that act as enhanced interfaces to your smart-phone. The smart watch market was started by a small Kickstarter project called Pebble a few years back, but has grown to include some large, international players such as Samsung and Nike. Apple has announced a smart watch for this spring. On the flip side, Google's revolutionary wearable computing project 'Glass' has actually been cancelled, perhaps due more to consumer concerns over privacy than a lack of hipsters keen to throw money at the project to get 'this year's cool toy'.

One thing in particular that makes these two technological advances interesting is that they share a common set of hardware requirements – low power, sensors and communications. Technology developed for one market addresses the other, so the technology convergence benefits both markets, and helps drive innovation and cost reduction.

Hobbyist IoT kit

While researching our next Kickstarter project (a PIC32 variant of the LPLC board) we stumbled across the British electronics company ByVac, at: **www.byvac.com**. They sell a variety of electronics development boards, but what caught our eye was the BV508 – a PIC32 processor on a board with a similar footprint to our LPLC board coupled with a tiny Wi-Fi transceiver board, at a very reasonable price. So cheap in fact that we abandoned our plans for a Kickstarter project and instead ordered a BV508 to discover whether this board could indeed be suitable for rapid prototyping of IoT projects.

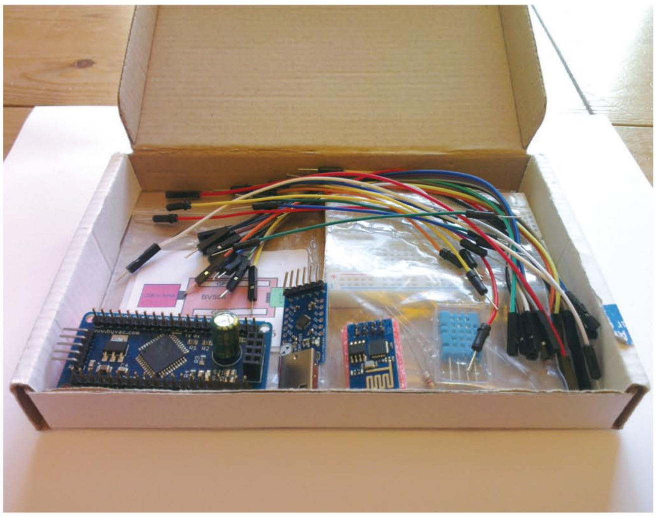


Fig.1 The BV508 IoT kit contents – a complete IoT development package

So what exactly is the BV508? It's a kit of parts for experimenting with Internet-connected projects. It consists of a PIC32 development board, equipped with SIL headers suitable for connecting it to a tiny PCB holding the Wi-Fi interface and SIL headers for connecting it to a breadboard. Their 'IoT Wi-Fi Discovery kit', for just £4 more, is a completely self-contained development kit adding a breadboard, USB-to-serial interface, combined temperature and humidity sensor and assorted hook-up cables and components to wire them all together. It's supplied in a small cardboard box that doubles as a handy storage case. I bring mine everywhere, setting it up in cafes and airport lounges when I have an hour or so spare. It does get the odd look! It really is a complete development kit – just add a PC. You can see the kit contents in Fig. 1.

The PIC32 processor is significantly more powerful than the PIC18F processor on the LPLC board; it's a 32-bit processor rather than our 8-bit one; it runs at 40MHz and has 256KB of flash memory on-board, plus a whopping 64KB of RAM. It draws more power than the PIC18F processor, but that's not a significant problem. IoT devices generally fall into one of two categories – those that can be powered from small batteries, and those that are powered from mains. Due to the use of a Wi-Fi interface this board falls into the latter classification – there is simply no way a coin cell could provide the instantaneous current required by this chip's radio transmitter circuit.

What makes this kit stand out (besides its low cost) is the inclusion of an intriguing Wi-Fi interface board based on a relatively new IC, the ESP8266.

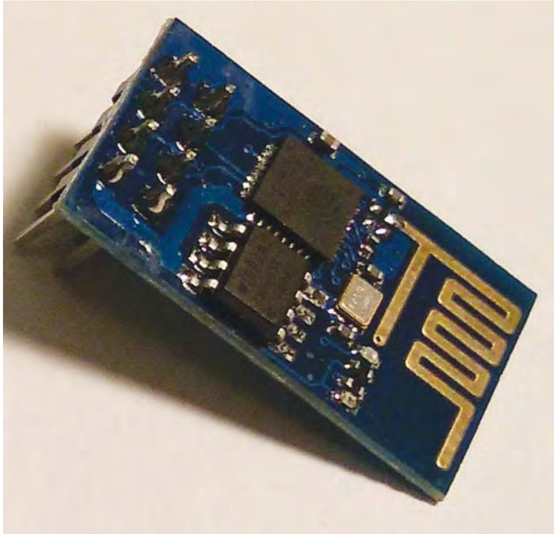


Fig.2 The ESP8266 board – a complete, stand-alone Wi-Fi interface

Wi-Fi interface

The ESP8266 is a fantastic device; it's a highly integrated 32bit CPU and Wi-Fi transceiver, all in a single integrated circuit. Boards based on this chip are tiny, and very, very cheap – testament to the highly integrated nature of the ESP8266. ByVac's offering is shown in Fig.2.

While technical details for this part were once difficult to find, there is now an active community developing applications for it, and there is a complete GNU GCC C compiler tool chain available for the embedded microcontroller. It's possible to create applications, which run directly on the IC itself, as the code is held in an external 8-pin flash memory IC. Developing applications for the chip itself does require significant embedded engineering experience, but that's where ByVac's product offering comes in – the PIC32 processor on the larger board is pre-programmed with the company's own BASIC-like programming language, which includes a library for accessing the ESP8266, and the example programs supplied with the board show how simple the applications themselves are. The chip contains an Xtensa LX106 32-bit processor (for which a full, open-source C compiler toolchain is available) and the radio transceiver operates in the 2.6GHz band, implementing 802.11 b/g/n protocols. Of particular interest is the ability of the hardware to support Access Point mode, meaning it is possible to broadcast it's presence, allowing mobile devices to connect to it, rather than the other way round. The demonstration application programmed into the processor uses this mode, hosting a simple website – all on a tiny PCB!

BASIC-like language

An introduction to the language would be an article in itself, but suffice to say, the language is very 'BASIC-like' but comes with some interesting features, such as libraries being compiled directly into machine code to increase the performance of user programs. The programming language is proprietary and currently unique to ByVac's products, they do, however, intend to make the language open source. For now, the language is freely available to program into your own PIC processors, something we may try out in the future. The language is very easy to learn and if you have any BASIC or C experience then you should be able to create your own applications within hours of getting your hands on a kit. We plan to put that statement to the test!

Setup and software installation

A simple A4 sheet of paper explains how to connect the boards together with the temperature and humidity sensor. Demonstration software pre-loaded on the PIC32 – the source code is available on the ByVac website – promises to provide a simple website for any devices connecting via Wi-Fi.

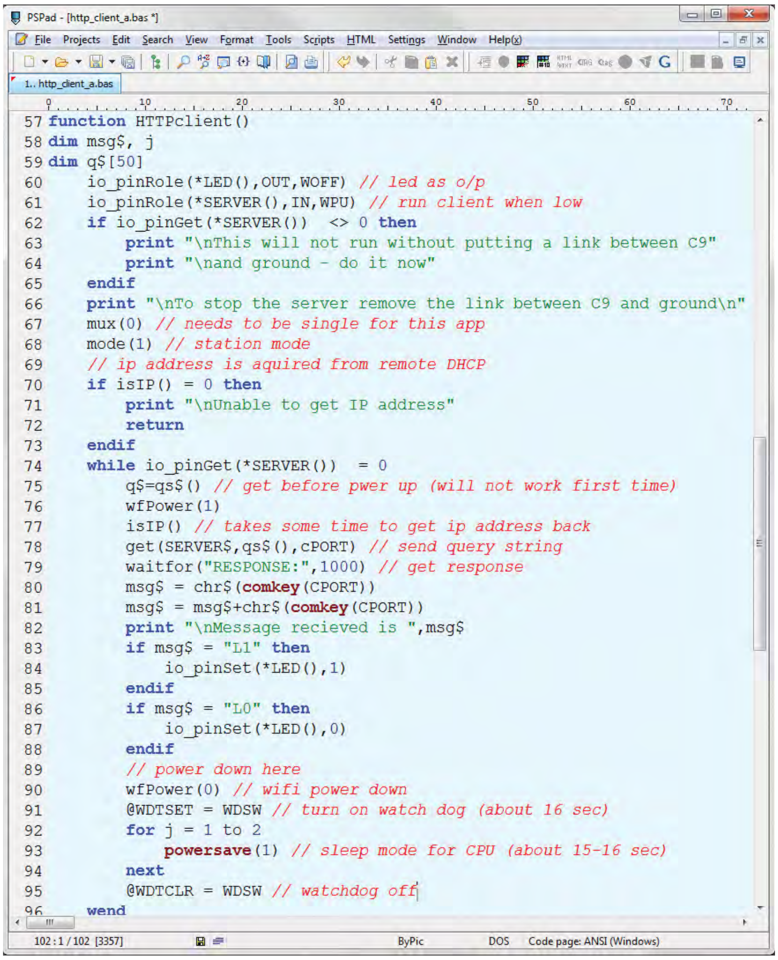


Fig.3 The sensor monitoring code is similar to BASIC, and very short due to the powerful built-in library functions

It took just 15 minutes to hook up the boards and connect to them via our mobile phone's Wi-Fi interface. The BV508 appears like any other Wi-Fi access point, providing open access. A very simple web page gives the option to read the current temperature, humidity or toggle the LED on and off.

Now, it's time to hunt down the code and modify it. We had to jump through a few hoops, cross-referencing examples found using Google, but the code to create alerts from the board (throughout our home Internet access point) was very simple, making use of one of the many cloud-based services designed to link IoT devices to the web – **pushingbox.com**. 'Pushingbox.com' is an on-line service that enables Internet-connected devices to push messages out to services such as email or Twitter, allowing you to create private or very public messages from your devices. It's a free service (they are probably still trying to work out how to monetise their service) and as a result the code to connect to them and broadcast the sensor is very simple – as easy as this:

```
function sendmessage(pushingbox_id$[30])

dim a$[160]

    a$= "/pushingbox?devid="+pushingbox_id
$+"&temperature=19&humidity=38"

    get("api.pushingbox.com",a$,80)

endf
```

We configured the pushingbox service to broadcast a twitter message once an hour, with a separate email message to ourselves when the temperature and humidity were outside of nominal limits.

Getting to this point took just three hours, and it's very exciting receiving automated twitter messages and emails from ones own hardware devices located around the home.

We had a bit of fun adding a range of responses and using a random number generator to choose between them, giving the message an almost Human quality. HAL it isn't, but fun it certainly is!

This has been a very brief, whistle-stop tour of the capabilities of the ByVac IoT Wi-Fi Discovery kit, but in just three hours we had loads of fun and quickly recognised the endless possibilities afforded by the hardware and the (unrelated) on-line services that are available, right now, to take messages from your hardware and forward them on as you see fit. It demonstrated how exciting this technology is, and how easy it is to start experimenting with it.

Power consumption

With the design as it stands, the current consumption is quite high – relatively speaking – drawing tens of milliamps even when 'inactive'. Although this figure sounds quite low, it is too high for a modest battery pack, but it is perfectly reasonable for use in a shed or workshop, running from a small, cheap wall-mounted PSU. Our board is now monitoring humidity and temperature in the *Pic n' Mix* workshop, which we have recently insulated, but want to ensure that in our cold wet winters the workshop remains a sensible place to store paints and tools. Now, should any of the monitored variables falls outside limits, the board will send a warning email with full details.

Power consumption is not a constant, however. Keywords exist within the BASIC programming language to place the processor in a low-power state, waking infrequently. The ESP8266 supports a number of low-power modes, and with careful further design work there is no reason why this circuit could not run off batteries for a reasonable length of time. The limiting factor will be the peak, instantaneous current required by the ESP8266 – up to 240mA. You are not going to be able to deliver that current with a coin cell, but it may be possible to create a design that could run from a set of 'C' or 'D' cells for a reasonable number of months. It all depends on your application; for ours, a plug-pack PSU is fine, and we can rest easy, forgetting about it until the next automated warning email comes in!

This article probably raises more questions than answers. 'Why pushingbox?', what is it, what other services are available? How do they make money from these services? Are they safe to use? We will answer those questions and more in upcoming articles, later this year.

Next month

In June's issue, we will take a look at improving the performance of our LPLC's LCD interface, and take a look at an intriguing PICKIT programmer interface that removes the need for a physical connector on your board designs.

Not all of Mike's technology tinkering and discussion makes it to print. You can follow the rest of it on Twitter at ***@MikeHibbett****, and from his blog at* ***mjhdesigns.com***

CIRCUIT SURGERY

REGULAR CLINIC

BY IAN BELL

Constant current sources – Part 3

FOR the last couple of months we have been looking at current sources prompted by a discussion on the *EPE Chat Zone* started by ***atferrari***. These articles introduced the concept of current sources and then mainly focused on op-amp-based circuits. The classic transistor current-mirror-based current source (Fig.1) is mentioned, but not discussed in much depth.

Recently, ***lost*** referred to these *Circuit Surgery* articles in a thread mainly centred on the discussion of transistor characteristics and *h*-parameters in the current *Teach-In 2015* series; ***lost*** wrote:

'I have another puzzle arising from Teach-In 2015 *(EPE, March 2015). Fig.2.4 (b) on page 39 and associated text indicates that an NPN transistor cannot supply (approx) constant current with V_{CE} below about 2V. In the same issue there is an article on constant current sources which show $V_{CE} = V_{BE}$ ie, about 0.6V. I understood that a transistor 'bottoms' when the base-collector junction becomes forward biased. With grounded emitter that shouldn't occur until V_{CE} is very low (tens of mV).'*

Later adding: *'It is clear that I really don't understand the source of the 'knee' on the output characteristic. That's a pretty basic thing not to know. I guess it must be buried in the physics rather than the electronics.'*

This month, we will therefore continue to look at current sources, specifically Fig.1, but extend our discussion to include the bipolar transistor output characteristics, how this relates to the output voltage range of the circuit in Fig.1 and to explain the knee in the characteristic.

lost's comment that $V_{CE} = V_{BE}$ refers to *Q1* not *Q2* in Fig.1. *Q1* is 'diode connected', so the transistor's base-emitter junction is effectively used as a diode. Given that V_{CC} and *R* are fixed, this means that the V_{BE} of *Q1* will be fixed – it is effectively at a constant voltage. This constant voltage is applied as the V_{BE} of *Q1* and in turn sets a constant collector current in *Q2*. Assuming that *Q1* and *Q2* are matched transistors (have identical characteristics) equal V_{BE} implies equal collector currents, so the current

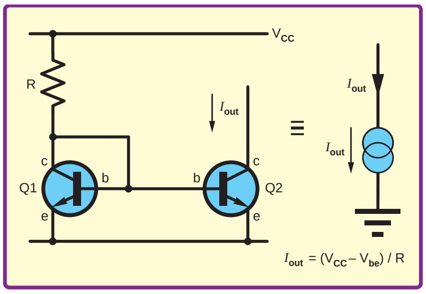


Fig.1. Constant current source (sink) based on a current mirror – Q1 and Q2 are matched transistors

source output current will be equal to the current in *R*, which is given by $(V_{CC} - V_{BE})/R$.

Source limitations

The circuit behaviour described in the previous paragraph is somewhat idealised and only applies over a limited set of conditions. I_{out} (I_C of *Q2*) will not remain completely constant when the collector-emitter voltage (V_{CE}) of *Q2* varies due to the less-than-infinite output resistance of *Q2*. Furthermore, *Q2* can only maintain the (nearly) constant current over a limited range of V_{CE} – this is the compliance range of the current source. At very high V_{CE} the transistor will break down, or burn up due to excessive power consumption. This is a relatively straightforward case of a voltage or power-handling limit being reached.

The lowest V_{CE} at which *Q2* maintains a constant current is a more subtle issue. As ***lost*** says, this is related to the base-collector junction of *Q2* becoming forward biased – *Q2* becomes saturated, or 'bottoms', as ***lost*** described it. Under these conditions V_{CE} remains more or less constant at typically 0.2 to 0.3V (*Q2's* V_{CEsat}) for quite a wide range of collector currents. The phrase 'wide range of collector currents' of course tells you that under these conditions the collector current is not acting as a constant current source.

Output characteristics

The figure from *Teach-In* that ***lost*** refers to, and which he tries to relate to the current source in Fig.1, is the output characteristics of an NPN transistor in common-emitter configuration. This is a set of plots of collector current (I_C) against collector-emitter voltage for a number of fixed base currents (I_B).

The output characteristics are plotted in some, but not all, transistor datasheets. We can also obtain a set of output characteristics by simulation in LTSpice. This comes with the usual caveat that results may be limited in accuracy by the model and parameters used, but they are sufficient for our purposes here. The schematic required is shown in Fig.2 and the results in Fig.3. The transistor is a 2N2222 – there is nothing special about this choice, it happens to be the first on the list of models provided by LTSpice. We use a DC sweep of two sources. Voltage source V1 sets the transistor's V_{CE} to different levels from 0V to 5V in 0.01V steps. This is done for each of seven values for I1, which sets the base current, from 0 to 600µA in 100µA steps.

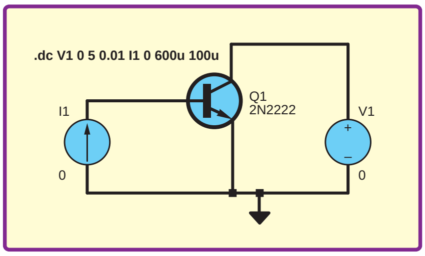


Fig.2. LTSpice schematic for plotting transistor output characterises

To a basic approximation, Fig.3 shows that for relatively large collector-emitter voltages, at a given base current, the collector current is constant (constant collector current implies a horizontal line on Fig.3). For example, at 300µA base current, the collector current is just over 50mA. If the collector-emitter voltage is decreased then there comes a point at which the constant current action stops – the 'knee' in the curves to which ***lost*** refers. Below the knee the collector-emitter voltage is almost constant at about 0.2V, for a wide range of collector currents and base currents (constant voltage implies a vertical line on Fig.3). At points below the knees in the characteristics in Fig.3 the transistor is in saturation mode and at points above this it is in forward active mode.

The above description was an approximation – the collector current is not completely constant in the forward active mode and the collector-emitter voltage is not completely

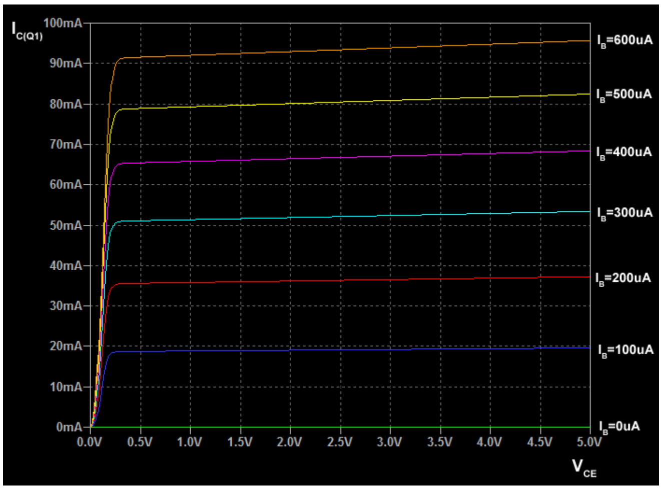


Fig.3. Transistor output characteristics obtained from LTSpice simulation of Fig.2

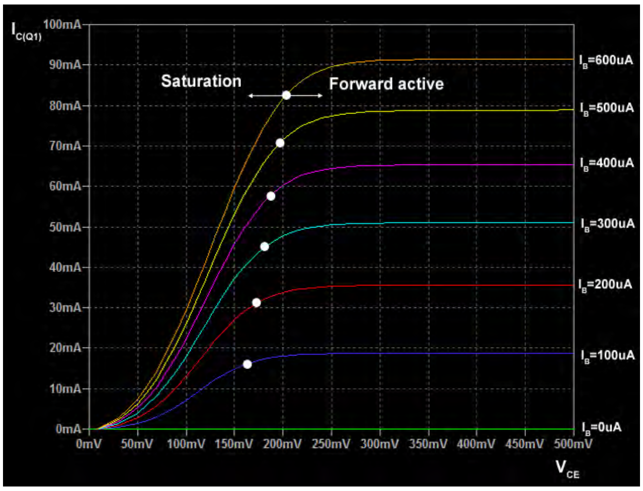


Fig.5. Transistor output characteristics. Zoom-in on Fig.3 for lower V_{CE} values. The dots indicate the transition point from forward active to saturation region of operation at 90% of the forward active current

constant in saturation. It is worth noting that when these characteristics are measured for real divides it is often done using pulses of applied voltage and current rather than continuous operation. This allows the characteristics to be obtained for values, which would damage the transistor, if applied continuously. The highest power dissipation occurs in the top right portion of Fig.3.

Looking first at 'constant' current operation we see that in forward active mode the collector current curves are approximately straight lines with small gradients. If we divide the collector current by the applied base current we get a figure for the transistor's forward current gain (known as the forward beta, β_F, which is equivalent to h_{FE} in the *h*-parameter model). For example, the base current of 100µA corresponds with about 20mA of collector current, giving a $\beta_F = I_C/I_B$ of 200. The current gain of a transistor in forward active mode is not completely constant – it varies to some extent with collector current; although for moderate changes in base and collector current it does not change by a large amount.

The gradient of a line on a current-against-voltage plot (as in Fig.3) has dimensions of 1/resistance. The shallow gradients in the constant-current region indicate a high resistance – this is the output resistance of the transistor operating as a current source (as in Fig.1). It is large, but not the infinite value of an ideal current source.

In Fig.3, we just about see the slope of the curves in the forward active region increasing at higher base currents. If we extend all these curves, as shown in Fig.4, the lines converge at a point far to the left on the V_{CE} axis. The value at which this occurs is called the Early Voltage (V_A) and is used to characterise this aspect of transistor behaviour – however, this is not the main concern of our current discussion, which is focused in saturation.

Saturation

Fig.5 shows a zoom in on Fig.3 for lower collector-emitter voltages. Here we can see that the transition from forward active mode to saturation mode is smooth – the knee in the curve is not sharp, so what point do we designate for the changeover? A typical approach is to use the point at which the collector current falls 90 per cent of its constant-current value (we can compensate for the Early effect if required to get an idealised value for this if necessary). Approximately 90 per cent of constant-current points are shown in Fig.5. The collector-emitter voltage at which this occurs is the saturation voltage (V_{CEsat}).

Looking at Fig.5, we see that the saturation voltage is not constant, so it should be quoted at specified collector and base current values. The Fairchild Semiconductor datasheet for the 2N2222 gives two values for maximum V_{CEsat}, firstly 0.3V at I_C = 150mA, I_B = 15mA; and secondly 1.0V at I_C = 500mA, I_B = 50mA. These currents are higher than those shown in Fig.3 and Fig.5, and reflect expected usage as a load switch.

In the saturation part of the curves in Fig.3 and Fig.5 we do not have a close-to-constant relationship between base current and collector current. The value of I_C/I_B is nowhere near constant for a transistor in saturation and is generally much lower than in the forward active region. In saturation mode we use the term 'forced beta' for the ratio of collector to base current, to distinguish it from the current gain in forward active mode.

The curves in Fig.5 between say V_{CE} = 50mV and V_{CE} = 150mV to 200mV (depending on base current) are more or less straight sloping lines – but they are not vertical. Thus, in the context of this graph, it is probably better to think of the transistor acting as a resistor in this region of operation, rather than having constant voltage across it. The idea that V_{CEsat} is a constant voltage makes more sense in a real circuit in which there is less extreme variation of conditions than in the set up in Fig.2, where V_{CE} is forced to take every value from 0V to 5V by the constant voltage source *V1*. A transistor in saturation can be modelled as a constant voltage of V_{CEsat} in series with a resistance, R_{CEsat}, if conditions do not change too much.

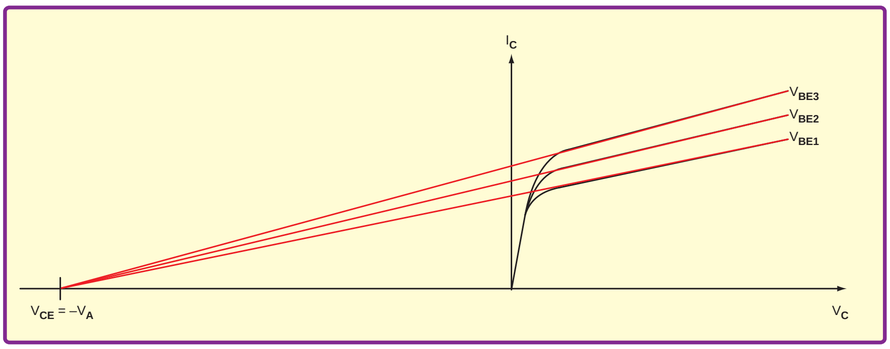


Fig.4. Finding the Early voltage from the transistor output characteristics

Switching

When a transistor is used as a switch it is typically switched between saturation mode and cut-off (not-conducting). The curve for I_B = 0 represents the cut-off condition in Fig.3 and Fig.5. The saturation voltage and effective resistance of the transistor in saturation is important when considering its switching performance. Higher saturation voltage and higher resistance result in increased power dissipation in the transistor when the load is on. At higher base currents V_{CEsat} is lower and

the slope of the characteristic curves in saturation is larger, so the resistance is lower. It follows that switching circuits often use relatively large base currents (compared to linear circuits).

To reduce power dissipation in switching circuits, semiconductor companies have developed transistors with low V_{CEsat} and saturation on resistance (R_{CEsat}), known as 'low V_{CEsat}' transistors. For example, NXP have a series of devices branded as 'breakthrough in small signal' (BISS) transistors which are marketed as having low power losses with higher energy efficiency than standard transistors in the same package. The PBSS4160PANPS device in this range has an NPN and a PNP transistor in an SMD package in which the NPN device has a V_{CEsat} of 120mV at I_C = 500mA, I_B = 50mA. The resistance from collector to emitter under these conditions is 0.24Ω. Compare the V_{CEsat} with that of 1.0V for the 2N2222 at the same current levels.

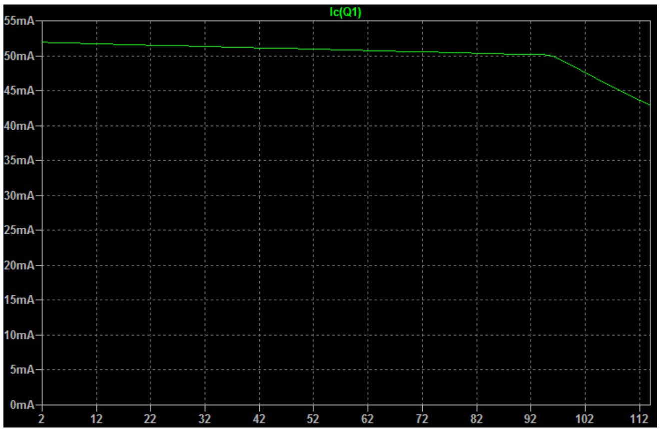


Fig.7. LTSpice plot of output current (IC(Q1)) against load resistance (R) for the current source circuit in Fig.6. Current is reasonably constant for loads up to about 94Ω

Back to the current source

The circuit in Fig.2 and the graphs in Fig.3 and Fig.5 go a long way to indicating what we might expect from the current source in Fig.1. However, the circuit in Fig.2 does not directly represent Fig.1. Therefore, the circuit in Fig.6 was used to simulate the circuit in Fig.1 in LTSpice with varying load resistance (R) connected between the current source output and the supply (5V). The simulation uses a parametric sweep of the resistor value (2Ω to 1000Ω) and calculates the DC operating point at each value.

The circuit was set up so that base current of Q1 would be about 300µA (hence an I_C of about 50mA), corresponding with one of the middle curves in Fig.3 and Fig.5. With a supply of 5V and a current of about 50mA we would expect the transistor to 'bottom out' for loads above about 100Ω (or just under) for which the voltage drop would be equal to or more than the supply at a current of 50mA.

The results from simulating the circuit in Fig.6 are shown in Fig.7 and Fig.8. Fig.7 shows that a near constant current is maintained for loads less than about 94Ω, as anticipated. Fig.8 shows that the constant current is maintained for V_{CE} values down around 200mV. The curve is similar to the 300µA curve in Fig.3.

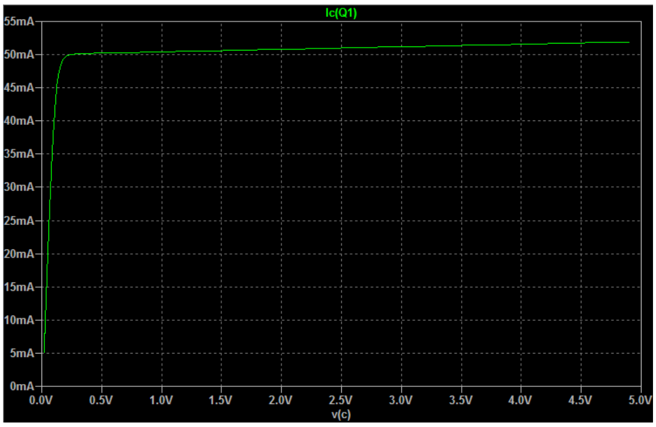


Fig.8. LTSpice plot showing output voltage compliance for the current source circuit in Fg.6 (output current against output voltage). The load resistance varies from 2Ω to 1000Ω. Current is reasonably constant for voltages down to about 200mV

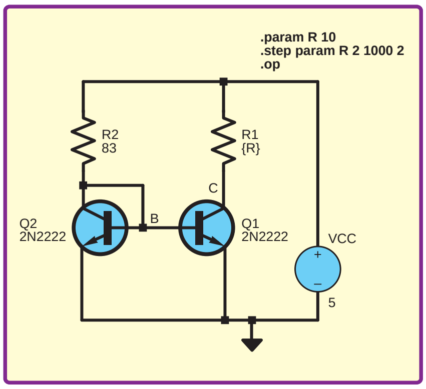


Fig.6. Circuit for current source simulation

Table 1: Bipolar transistor regions of operation

Base-emitter junction bias	Base-collector junction bias	Region of operation	Comment
Forward	Forward	Saturation	Switch ON
Forward	Reverse	Reverse Active	Poor amplifier, specialist uses
Reverse	Forward	Forward Active	Good amplifier
Reverse	Reverse	Cutoff	Switch OFF

Understanding the knee

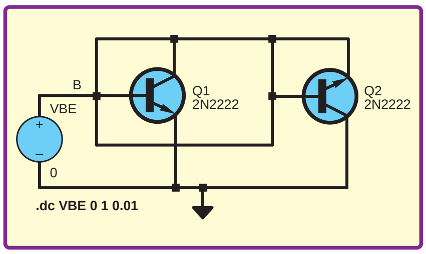
Having seen the difference between saturation and forward active operation from the perspective of the characteristic curves, we will shift to looking at it in terms of internal operation of the transistor. This will provide an explanation for the knee in the output characteristic – the transition between forward active and saturation mode.

A bipolar junction transistor contains two diode (PN) junctions, this is true of both NPN and PNP transistors; the junctions are simply connected the opposite way in the two types, requiring opposite voltages for the same operation. The two junctions can be either forward biased (on) or reverse biased (off) so there are actually four different regions of operation for the transistor, see Table 1 – note that we have mentioned three of these so far. The fourth is the reverse active region, which is not commonly used, but does have some applications in some types of logic and analogue switching circuits.

Reverse operation is obtained by swapping the collector and emitter connections. A transistor is semi-symmetrical, so it will still work as a transistor, but with very poor performance in conventional terms (eg, very low gain). If a transistor operating in reverse active mode 'bottoms' it will enter saturation just as a transistor in forward active mode does. Rememeber that the two PN junctions in a transistor are not identical. This is because the physical structure of the transistor is not fully symmetrical and the doping of the emitter and collector are not the same. We can investigate this by looking at the voltage-current characteristics of the two junctions.

Fig.9 shows an LTSpice schematic for plotting the characteristics of the base-emitter and base-collector junctions of a transistor. The two copies of the transistor are diode connected (as is *Q1* in Fig.1); this shorts out the other junction so we can see the change of current with applied voltage for just the junction of interest. The results are shown

Fig.9. LTSpice circuit to plot characteristics of base-emitter and base-collector junctions



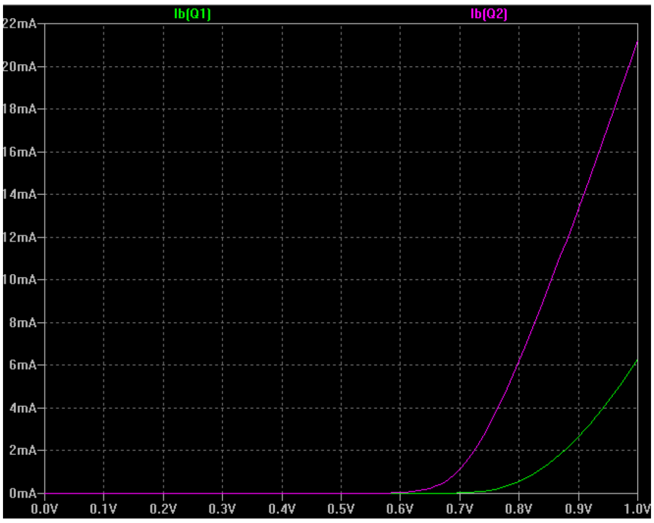


Fig.10. Plot of base-emitter (Q1, green trace) and base-collector (Q2, magenta trace) junction characteristics obtained from the circuit in Fig.9. The base-collector junction conducts at a lower voltage than the base-emitter junction

in Fig.10, from which we see that the base-collector junction conducts at a lower voltage than the base-emitter junction.

The transistor is more than just two independent PN junctions – it is the interaction between these junctions that gives us transistor action. In forward active mode the base-collector junction is reverse biased and the base-emitter junction is forward biased and conducts. In an NPN transistor this involves electrons moving from the emitter to the base (they are 'emitted' by the emitter). If all we had was an isolated PN junction, then all this current would flow through the base connection, but because of the voltage on the collector, the thinness of the base region and the relative doping levels most of these electrons flow to the collector (they are 'collected' by it). This current is through a reverse-biased junction – which occurs because the electrons are minority carries from the perspective of the P side of the junction and are able to cross it (electrons from the collector and holes from the base are prevented from flowing by the reverse bias, but electrons in the base are free to move across the junction).

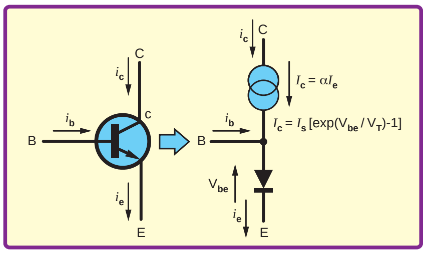
Transistor models

The previous discussion leads to a model of the transistor shown in Fig.11. The base-emitter junction is represented as a diode as we might expect. Its voltage-current relationship is given by the diode equation shown in the figure. The exponential relationship can be seen in the graphs in Fig.10. The collector current is modelled as a current source for which the current is given by αI_E. The term 'α' is the proportion of emitter current that carries on to the collector, avoiding the base. A typical value is 0.99. Applying Kirchhoff's current law we find that $I_B = I_E - \alpha I_E$, so if α is 0.99, 1% of the emitter current flows through the base connection and 99% through the collector.

The model in Fig.11 is not able to explain saturation and the knee in the output characteristics. For this we have to include the reverse behaviour of the transistor. This is like Fig.11, but the opposite way round (just swap the emitter and collector labels in Fig.11). Such a reversed model with a collector diode and emitter current source would cover just reverse active mode. It differs from the forward model only in different diode equation parameters to account for the different characteristics of the base-collector junction and a different value of α. The value of α for reverse mode is much less close to 1 than for the forward active case (typically 0.5 rather than 0.99), which accounts for the poor transistor performance in reverse mode.

We can combine the forward and reverse versions of the equivalent circuit in Fig.11 to produce the circuit shown in Fig.12. This is the classic Ebers-Moll model of the bipolar transistor. With reference to this model, consider an NPN transistor in forward active mode. The base will be sufficiently positive with respect to the emitter (about 0.7V) for a reasonable

Fig.11. Simple model of an NPN transistor without "reverse" behaviour. This covers forward active mode, but not saturation and reverse active



base-emitter junction current to flow (I_F in Fig.12), most of this current (αI_F) will flow via the collector and a small amount ((1 – α)I_F) via the base. The collector voltage will be sufficiently positive for the base-collector junction to be reverse biased, so I_R will be zero and therefore so will αI_R. The model could be simplified to Fig.11.

Saturation in Ebers Moll

Now assume the collector voltage is gradually reduced. Eventually it will drop below the base voltage, so the base will become positive with respect to the collector. Once this positive collector-to-base voltage is large enough, the junction will conduct (as current I_R in Fig.12). Under these conditions, we see from Fig.12 (using Kirchhoff's current law) that $I_C = \alpha_F I_E - I_R$. Substituting the diode equations for the two junctions into this equation we get:

In which V_T is the 'thermal voltage', a common parameter in semiconductor models, which is constant at a given temperature.

If we assume that V_{BE} is constant and around 0.7V then the first exponential term will be constant and positive – this term gives us the constant forward active collector current for a given V_{BE} at large V_{CE}. However, the base-collector junction is also conducting (in saturation) so the second exponential term will be positive and controlled by the value of V_{BC}. This term subtracts from the constant first term and so reduces I_C as V_{BC} increases. Decreasing V_{CE} makes V_{BC} larger and so decreases I_C. This accounts for the knee (decreasing I_C as V_{CE} is lowered) which we see in the output characteristic.

The point at which the knee starts is the point at which the base-collector junction starts to conduct. We see from Fig.10 that for the simulated 2N2222 this is about 0.5V, so saturation will start occurring for base-collector voltages above this value. If the transistor is just conducting then V_{BE} is about 0.7V according to Fig.10. If V_{BE} = 0.7V and V_{BC} = 0.5 then V_{CE} will be 0.7 – 0.5V = 0.2V, so we can say the saturation voltage is about 0.2V, which is consistent with Fig.5. For higher base currents we need higher V_{BE}, so we would expect the saturation voltage ($V_{BE} - V_{BC}$ at V_{CE} turn on) to be higher too, which is what we see in Fig.5.

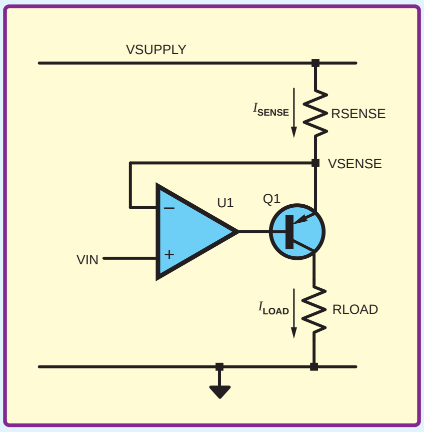


Fig.8. Current source with grounded load

PLEASE TAKE NOTE

Ian Bell, Circuit Surgery's author, has asked us to correct an error in Fig.8 of last month's column (April 2015). In the schematic, the op amp's non-inverting input should be labelled 'VIN'.

READOUT

Matt Pulzer addresses some of the general points readers have raised. Have you anything interesting to say? Drop us a line!

WIN AN ATLAS LCR ANALYSER WORTH £79

An Atlas LCR Passive Component Analyser, kindly donated by Peak Electronic Design Ltd, will be awarded to the author of the Letter Of The Month. The Atlas LCR automatically measures inductance from 1mH to 10H, capacitance from 1pF to 10,000μF and resistance from 1Ω to 2MΩ with a basic accuracy of 1%. www.peakelec.co.uk

All letters quoted here have previously been replied to directly

Email: editorial@wimborne.co.uk

★ LETTER OF THE MONTH ★

PCB design and production

Dear editor
A comment on Mike Hibbett's *Make Your Own PCB* series (*EPE*, September – December 2014). A few years ago I came across a program called Sprint-Layout 5.0 (now updated to 6.0) by a German company called Abacom, see: **www.abacom-online.de/uk**. So far, this excellent program does not seem to have had a mention, and it costs less than £40 to download. (Or did!)

It is remarkably comprehensive. Although it is limited to two layers, it can output Gerber, CNC drilling, and screen printing files as well as normal print output. It allows measurements in mils or mm, (though the grid stays as mil spacing). Also, you can import a bitmap file to use as a background guide to copy an existing layout, and it comes with an astonishing array of ready-made components, to which you can add your own component make-ups, or store whole sections of a layout as a 'component'.

Abacom complement Sprint with other programs for schematic and panel design, plus an excellent tone generator.

One other tip that readers might find useful – sunshine can be used as a UV source to produce PCB prints on photo-sensitive board.

You need a layout printed on transparent plastic and clamped to the board between a stiff backing board and a suitable piece of Perspex (or even glass). It will produce a very good result when held or placed facing the sun for about 100-120 seconds. This compares with about six to seven minutes in a UV box. It even works in bright cloud, but with a longer exposure.

Lloyd Stickells, by email

Matt Pulzer replies:
Thank you Lloyd – there are many excellent design packages available, but sadly we can't review all of them. However, it is always handy to have guidance and experienced reader feedback.

Not just for Windows

Dear editor
Your advertising for *EPE* in 'PDF format' (an acronym tautology, like 'PIN number') indicates that these files will require 'a Windows PC using Adobe Acrobat reader'. Now, hang on a mo' ... I know that many readers (and potential readers) of your excellent magazine are now using other platforms such as Linux and Apple OS X.

I have no problems opening PDFs on my two Linux Ubuntu PCs, so why should I need Microsoft and Adobe to enjoy your CD-ROMs? I'm sincerely concerned that you may be limiting your market, or failing to recognise the impact that open source systems are having in the world of personal computing.

Please consult with Alan Winstanley on this. Perhaps he can test your CD-ROMs on some other kit and mention the results in one of his excellent articles.

Pete Barrett, Northumberland, UK

Matt Pulzer replies:
Yup – guilty as charged! Not only did we print a tautology, but spurned our loyal Linux/Apple readers – ironic, given that I am typing this on my Mac Pro.

I think almost every PC under the sun can now read PDF files, but the encryption used on our PDFs is very sensitive to operating systems, ie, on one system it will work fine, but if you were to try to open on another system that is using a certain firewall, for example, the PDF will not open. If you would like to trial the encrypted PDF to see if it does work on your operating system then please email stewart.kearn@wimborne.co.uk.

Panel confusion

Dear editor
I am making the *SidRadio* project and considering whether to buy the front and rear panels. However, the dimensions given in the magazine for these panels seem small for the recommended case size.

The article in the October 2014 magazine gives the size of the panels as 200 × 30mm, for a recommended case sized 225 × 165 × 40mm.

I managed to obtain an ABS case of this size from Farnell (1526700 G747A case, ABS, Alu panel, 225 × 165 × 40mm), which looks practically identical to the photographs in the magazine. However, the metal end panels that come with this case are sized 219 × 34mm. This means that your end panels if they are really 200 × 30mm would be far too small and would not be retained by the rebates in the case.

Phil Turrell, Basingstoke

Matt Pulzer replies:
You're quite correct – the dimensions given in the parts list are wrong. The panel sizes in both our drawing and PCB layout programs are 219 × 34.5mm and indeed the panels that we sell are that size.

Transistor bank

Dear editor
That's an interesting collection of historic transistors on the front cover of the February *EPE*. I couldn't identify them all, but a Russian one stood out. Here's a photo of the similar MP41 in my own collection (our 'P' looks like a capital Greek letter 'Pi' in Cyrillic). It was probably made in the mid-1960s, this one still works.

You might like to print a mention of the Transistor Bank that I offer. If a reader has genuine need of an old device, perhaps to restore vintage equipment, then I am happy to be contacted. Should I have the required transistor or diode (or a suitable substitute) available, I'll send it on as long as a pre-paid self-addressed envelope (of adequate

size) is supplied. The range is obviously limited, no promises, but enquiries are welcome: **cgmm2@btinternet.com**. 63 The Drive, Edgeware, Middlesex, HA8 8PS

Godfrey Manning, G4GLM

Matt Pulzer replies:
I must confess the transistors chosen for the Teach-In *cover were chosen simply for their 'artistic merit' rather than their semiconductor behaviour.*

Thank you for your very generous Transistor Bank offer. I'm sure readers will be in touch.

EE Beta: a blast from the past

Dear editor
I've been trying to get hold of back issues or articles on the *Everyday Electronics Beta Guitar* project from issues November 1972 to January 1973. The three articles covered the construction of the guitar, including a blueprint. I actually built it back in 1973 at woodwork class and used it in our local band for a couple of years before buying a guitar. I still have the *EE Beta*!

I'd like to repair the circuitry and show my kids the articles from which it came. Do you know where I can get them? Hope you can help.

Fred, Australia

Alan Winstanley replies:
I asked for help from a friend of EPE *known to many on our Chat Zone as 'magman' who collects magazines as a hobby. As this identical question had cropped up before, we put Fred in direct contact and although Mike can't promise to assist with every enquiry he was happy to help with a scan of these particular back issues at our request. In fact, thanks to the Internet the scans were there for Fred's breakfast-time.*

Fred replied:
Wow…! did these things bring back memories. I have always been into electronics and *Everyday Electronics* was one of my favourite mags. The guitar came to me by a lucky combination of circumstances. I was in secondary school and subjects included woodwork. In my fourth year (1973) we had the opportunity to pick our own project to build. What a coincidence that *EE* just happened to come out with the *Beta Guitar* project. So I built it, learned to play and got together with some mates and had a band for a while. I did buy a guitar after a couple of years, but always kept the *EE Beta*.

I have a few guitars these days, but when I recently got together with some of the old boys in the band, we talked about the old music days and of course that big, red, loud guitar that I had built. I plan to resurrect it and need to rebuild the electronics, thus the need for the articles. I also wanted to let the kids see how this thing came to be. I'm hoping to get them hooked on electronics and/or music. Well thanks once again to everybody, from me – Fred Treven – 'down under' in Melbourne, Australia!

'k' for konfusion

Dear editor
It interests me that in your articles you use the prefix 'k' for kilo. I was always taught throughout school and university (albeit a long, long time ago!) that any prefix larger than the unit must be in upper case and smaller in the lower case: ie, Mm is megametre and mm is millimetre, likewise Dm is decametre, whereas dm would be decimetere – hence, kilometre should be Km.

Alister Bottomley, Strathblane

Matt Pulzer replies:
I'm afraid you have been mislead! A quick review of the following will confirm this:

http://physics.nist.gov/cuu/Units/prefixes.html
http://en.wikipedia.org/wiki/Metric_prefix

While I certainly agree that it would be a neat and handy scheme to arrange prefixes in the way you discuss, unfortunately it is not the (current) agreed approach – kilo, (as well as hecto and deca) are all lower case – k, h, da.

I say 'current' because the system is updated and refined from time to time. Prior to 1960, the little-used deca prefix had a confusing variety of prefixes: dk, D and Da. My Austrian grandmother had cooking weights in decagrams, all marked in 'Dg'. Nowadays, however, decagramophiles would use 'dag'. Perhaps this is where the confusion arose?

To the best of my knowledge, hecto and kilo have always been indicated with 'k' and 'h'. This is especially important for kilo because uppercase 'k' is reserved for 'kelvin', the fundamental unit of temperature – named for the great physicist and engineer Lord Kelvin, who was Professor of Natural Philosophy for over 50 years at Glasgow University, a stone's throw or two from your home.

Note that this is a pedant's nightmare. The man is Kelvin (uppercase), but the unit is kelvin (lower case), indicated with 'K' (upper case). A thousand kelvins, or a kilokelvin would be 1kK (upper and lower case)!

One other source of confusion, particularly for anyone working in electronics, is the kilobyte. A kilobyte is not 1000 bytes, but 1024 bytes (2^10). To indicate this numerical difference upper case 'K' is used – two kilobytes is written as '2KB'

I discussed this with EPE's Cool Beans *columnist Max, who declared: 'the whole k vs K thing is a pain in the rear end – I understand why they originally decided to use a lowercase 'k' – but I wish they hadn't'.*

He asked me if I had 'seen the stuff about using kibi (Ki), mebi (Mi), gibi (Gi) qualifiers for computer memory?'

I hadn't, and now pass this on to you for even more to remember:

http://en.wikipedia.org/wiki/Binary_prefix

Who'd have thought a thousand could be so complicated?!

Cold caller revenge… *EPE* style

Dear editor
I've been an avid reader of *EPE* for many years. It's a great source of technical info, but people have stopped inviting me to parties.

I even seem to be useless at phone conversations. Please can you help me? Here's an example from last week...

Pete Barrett, Northumberland

Fred: *Hi, this is Fred. I'm calling to make you aware that you're eligible for a government 'green energy' grant to install solar panels.*
Me: *Hi Fred, thanks for calling. Tell me, do you use the first, second, third or fourth generation solar collectors?*
Fred: *Errr ... they're solar cells that make electricity.*
Me: *All those I mentioned are photo-voltaic cells, that's just the generic name. Let's start over. Are they thick-film silicon p-n junction, amorphous silicon, cadmium-telluride, gallium-arsenide or optical rectenna systems? Also, are they lensed or un-lensed collectors?*
Fred: *You seem to be quite clued-up on this subject.*
Me: *Yeah, I'm an EPE reader, it's full of great articles explaining recent developments in electronics and…*
Fred: *CLICK… dial tone*
Me: *Nice chap, but maybe he thinks I'm after his job.*

Matt Pulzer replies:
We think you are doing just fine, and you definitely don't need our help!

AUDIO OUT

By Jake Rothman

Ge-mania – Part 2

Hi-Fi history

A company that epitomises the history of amplifier design from valve to silicon is HJ Leak and Co Ltd. It no longer exists, but the brand remains, owned by Stan Curtis's company International Audio Group. The Leak Stereo 30 was the first successful production transistor amplifier in the UK, a 'productionised' version of a Toby Dinsdale design, released in early 1964. The circuit (Fig.14) used relatively high-voltage transistors (for germanium) the NPN AC127Z, along with a high frequency video AF118 transistor for the VAS stage. Output transistors were AD140s rated at V_{ce} = 55V and I_c = 3A (costing a pricey £1.25 each in 1964, roughly £22.50 in today's money) giving 12W into 15Ω, a similar output to the valve Stereo 20 it replaced. Thermistors were necessary for stabilising the DC mid-point and quiescent current. The distortion was worse than the valve amplifier and harmonic plots are given in Stephen Spicer's book, *Firsts in High Fidelity: The Products and History of H.J Leak & Co. LTD.* Later in 1968, the Stereo 30 was converted to silicon, becoming the 35W Stereo 70, and the improved stability allowed the thermistors to be discarded. The noise and distortion were also significantly reduced. A 20W version, the Stereo 30 Plus was introduced later. Uniquely, there was a resistor from V+ to put 2mA into the base resistor of the lower output transistor in Leak's silicon amplifiers. This eliminated sharp switching spikes and the 1.8kΩ resistor in the germanium circuit may also serve a similar function. In 1971 the front panels were changed to thick aluminium to make the Delta 70 and 30, which were in production until 1974 (Fig.15).

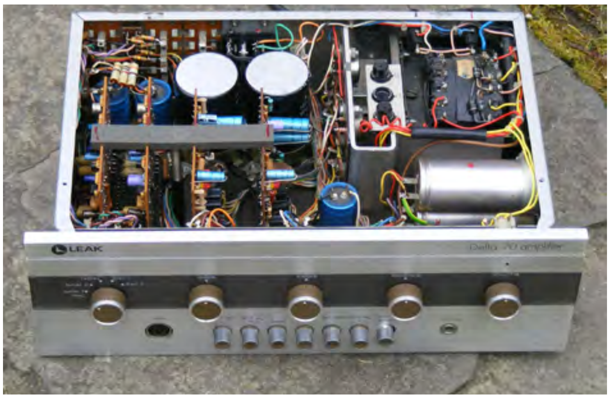


Fig.15. Leak Delta 70, a popular quasi-complementary silicon amplifier. It was a direct descendent of the germanium Stereo 30. This example was made in 1974 and still sounds excellent

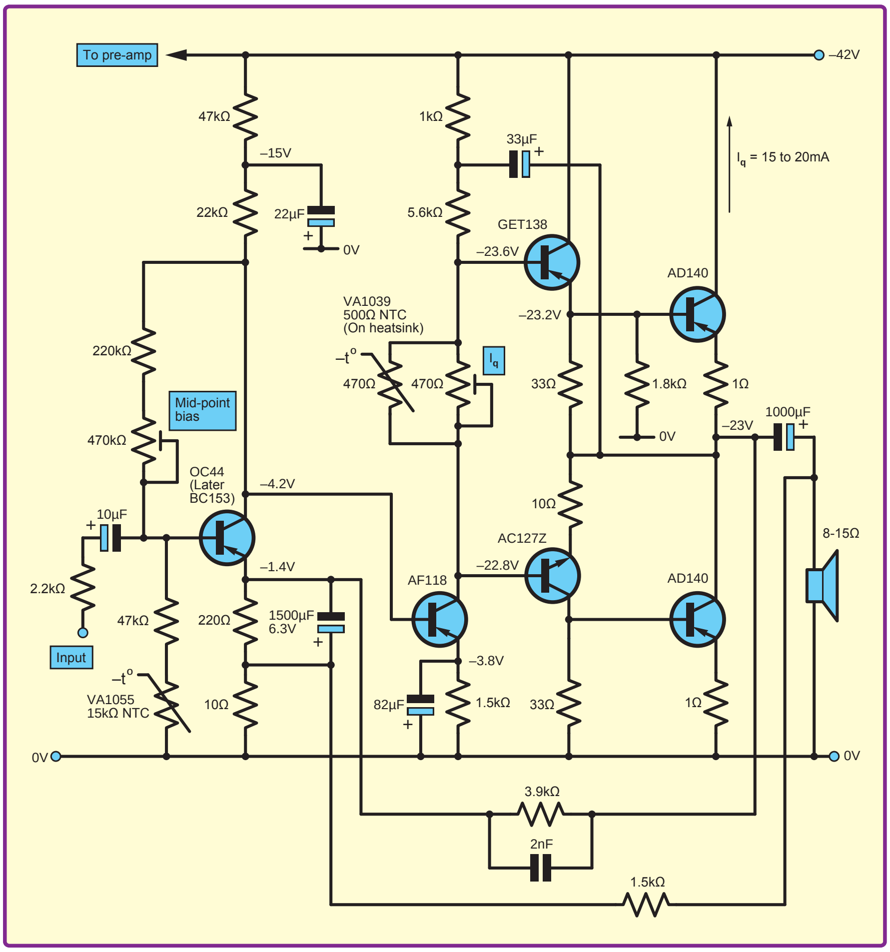


Fig.14. Leak Stereo 30 circuit; the UK's first commercially successful transistor Hi-Fi amplifier

'Transistor sound'

Other amplifier designs were not so fortunate, the shortcomings of the

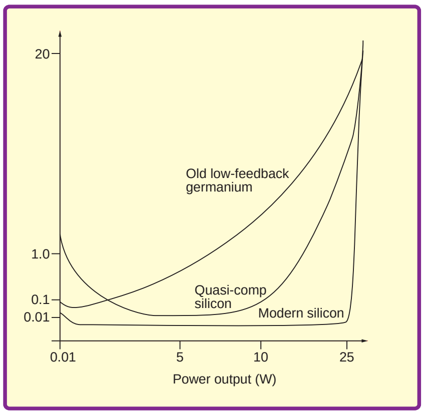


Fig.16. Comparison of distortion vs output curves of early germanium, silicon quasi-complementary and modern high feedback silicon power amplifiers

Fig.18. Bush 1961 germanium radio

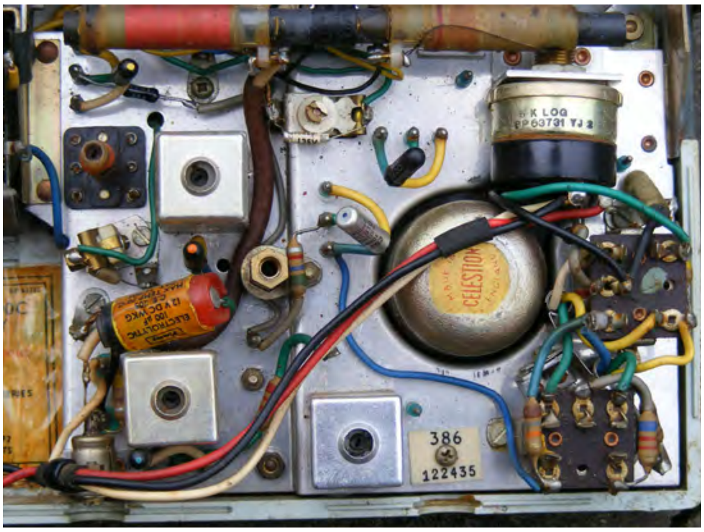


Fig.19. Bush radio internal construction; note the hard-wired chassis construction and large number of transformers

quasi-complementary output stage becoming much more apparent with silicon, and the 3× wider asymmetrical crossover region gave rise to the public perception of an unpleasant ‘transistor sound’. This problem was exacerbated by the typical use of bootstrapped collector loads on the driver stage, which were fed from the output stage. The output stage lacked gain at low levels, just where the boost was needed. Baxandall developed a diode dodge to fix the output asymmetry, but the damage to the audio reputation of the transistor still persists. Cyrus and Naim went on to use the Baxandall diode, along with a constant current driver load, with great success for many years. The all-silicon Quad 303 was the first to fully banish ‘transistor sound’ in 1967 with a complex triple-based output. Germanium’s brief foray into the world of Hi-Fi was over.

The use of fully complementary outputs, when silicon PNP transistors became available at a reasonable price in the late 1970s, effectively fixed the crossover distortion problem for most listeners. Using a long-tailed pair input stage and constant current VAS loading, this enhanced version of the Lin topology, popularised by Douglas Self *et al*, remains the mainstay of amplifier design today. The Leak 2000 series introduced in 1974 were of this type, and like many used dual power rails, eliminating the bulky output capacitor. Little has changed since, apart from the introduction of high-speed MOSFETs allowing PWM or class D (not digital!) to take over where high efficiency and light weight is needed. Crossover distortion is not an issue but there are other linearity, switching and EMI problems.

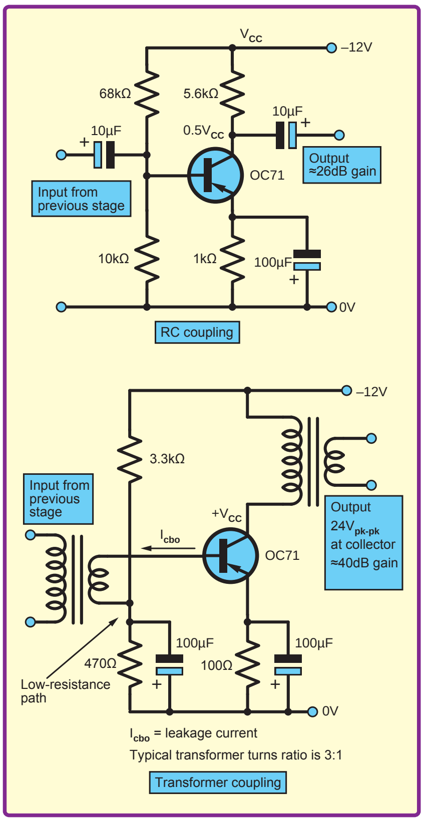


Fig.17. Transformer coupling vs resistor-capacitor (RC) coupling

Musical market

So much for audiophiles – what musicians want is different to Hi-Fi enthusiasts. They typically look for gradual distortion build-up with lots of second harmonic, akin to a soundboard being hit or a string being plucked hard. Modern amplifier technology with its large feedback factor doesn’t do this, it’s dead clean until it clips very hard – an unnatural effect (Fig.16). The ‘sound’ of germanium equipment is mostly due to the early low-feedback circuit topologies used, rather than germanium *per se*. Transistors were expensive, so the absolute minimum number were used, giving low open-loop gain.

Joule thief

To maximise gain and deal with the high leakage currents (due to the higher intrinsic conductivity of germanium) transformer coupling was often employed. This doubles (at least) voltage efficiency, since the full rail voltage is available at the collector rather than just half the rail voltage with resistor capacitor (RC) coupling (Fig.17). Inductive coupling and a low turn-on voltage also have the benefit of providing fantastic

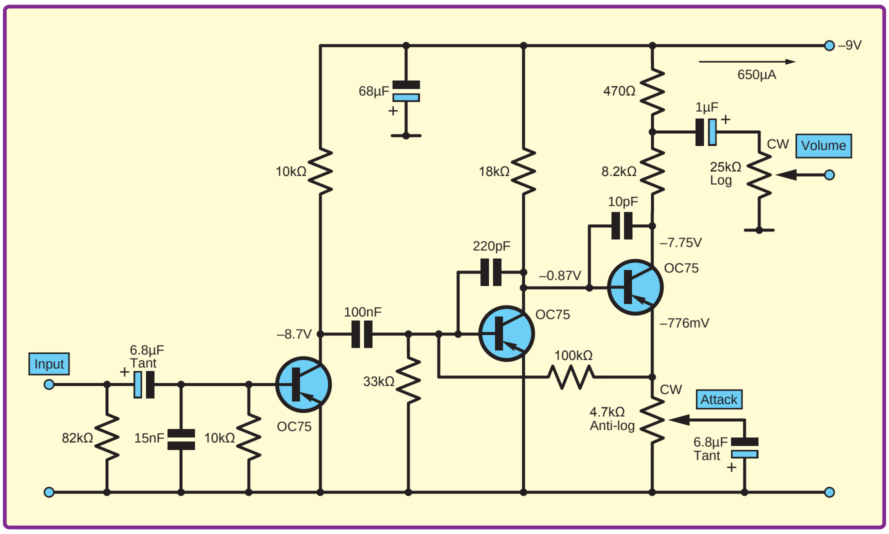


Fig.20. Circuit for the Sola Sound Pro Mk II fuzz box, a 1965 germanium design still in production, which gives that ultra-distorted Jimmy Hendrix sound. (Copyright Macaris, owners of Colorsound and Sola Sound brands)

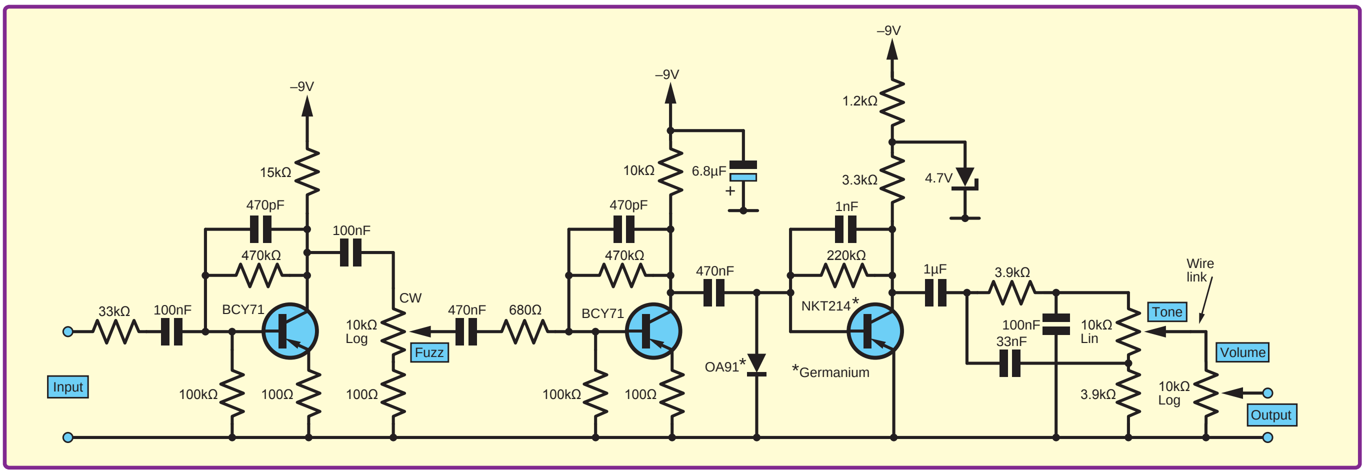


Fig.21. Circuit for the Colorsound Germanium Tonebender; note the germanium output transistor with diode to stop bias shift

battery life. My old Bush 1961 radio (see Figs 18 and 19) uses just 15mA from a PP7 9V battery, which lasts two years and works down to 3V. Such circuitry that squeezes the last juice from the battery is sometimes described as a 'joule thief', an Internet term. It's often used for LED flashers driven by a single AA cell, where germanium transistors give the best results. Tharma, Mullard's audio amplifier applications designer in the late 1960s, said that a 10% reduction in bias voltage in a germanium transistor changes the collector current (I_q) by only two times. By contrast, with silicon, the same percentage change will cause a 15-fold change in I_q. These factors explain why my 1978 ITT silicon discrete radio uses 35mA and works to 6.5V. My chip-based 1983 Grundig uses 80mA, but stops operating at 7V and the new DAB radio uses an astonishing 250mA, with the battery going flat after a few days use. This is a bizarre trend, given our current commitment to energy reduction – and all just to listen to Radio 4. It amazes me that broadcasters in their now abandoned digital radio switchover have transferred a huge amount of the electricity bill from their transmitters to their listeners without anyone noticing!

Dealing with leakage

When early transistors got cheaper, transformer coupling gradually gave way to RC coupling with low resistance potential dividers for bias. Occasionally, direct coupling was used, but this had a tendency to drift. I make a 1965-design fuzz box, the Sola Sound Pro Mk II (Fig.20) that sounds wonderfully 'germaniumish', but if you leave it in a sunny shop window, it can bias itself off. The first stage is self-biased and it has to be selected. Indeed, all germanium devices have to be individually sorted and tested before use because leakage, noise and gain vary much more than modern devices. If you get one that measures an H_{fe} of 500 on your multi-meter don't believe it; that's the leakage current making it look better than it is. About 80 is considered good, however some testers such as the Peak DAC55, also give a leakage

Fig.22. Colorsound Germanium Tonebender in 1970s colour scheme; note the metal-can NKT214F germanium output transistor

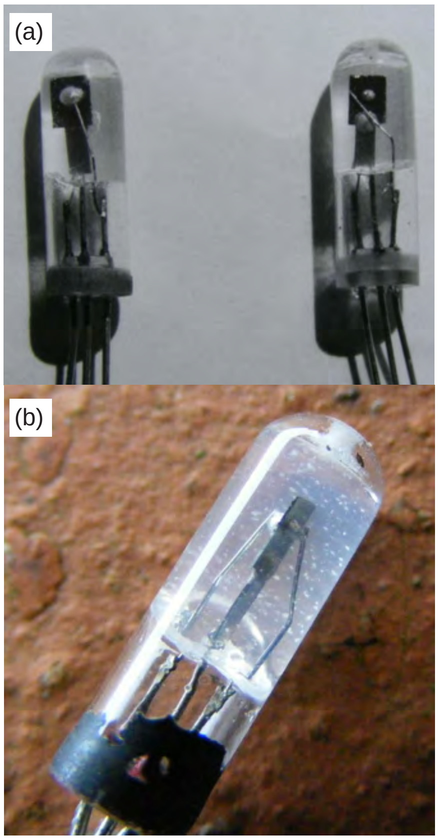


Fig.23 (Top) Old germanium glass-cased transistors; notice the disconnected wire. (Bottom) OC71 with paint scraped off – a popular dodge in the 1960s to get a cheap photo transistor

Fig.24. OC42 with metal sleeve removed to show glass case with opaque filling

current figure. It does this by applying 5V across the collector and emitter with V_{be} at 0V. A figure of more than 0.5mA is considered unacceptable. The Peak analyser takes leakage into account to give an accurate H_{fe} figure. If the base lead is left floating, or fed from a high resistance source, the leakage current is much worse than if it is grounded. This is why transformer coupling works especially well with germanium transistors since the low DC resistance of the transformer minimises the effect of the leakage current. Leakage current is generally noisy and Tim Orr used this for white noise in his synthesiser designs, he also used to 'cook' his transistors with an excessive current pulse to make them even noisier.

Using only a pull-down resistor, typically 10kΩ, is a technique often used in fuzz boxes to provide a form of self-bias, using just leakage current. As expected, leakage current increases with temperature. One method to compensate for this is to wire a reverse-biased germanium diode from the base to ground, or better still, to a reverse potential equal to the collector voltage. Such diodes can also prevent charge build-up on input coupling capacitors. This can cause the transistor to become biased off when overdriven. This technique is used in the *Germanium Tonebender* fuzz box I designed for Colorsound – see Fig.21 and 22. A strange quirk of germanium transistors is their reverse V_{be} rating is around 10V, as opposed to the 5V of planar silicon transistors. This is not normally an issue, but it occasionally causes shorter time constants in multivibrators if the transistors were changed to silicon versions.

Germanium reliability

Over 50 years of service have shown the hermetic SO-2 glass-cased devices like the black-painted OC71 and metal-sleeved OC/42/72/81 are much more likely to work than metal-cased types, which are affected by internal tin whisker dendrite growth shorts, poor sealing and internal solder flux contamination. Germanium devices need hermetic cases, so can't use epoxy cases. There's a good NASA site (**http://nepp.nasa.gov/whisker**) on the 'tin problem' with internal electron microscope and X-ray images of AF114 RF transistors. Another problem is the mechanical fragility of germanium devices. Occasionally, the alloyed contacts would just fall off, as shown in Fig.23a. These early OC71 transistors were originally filled with clear silicone grease. Often these types were used as photo transistors with the paint scraped off (Fig.23b). Later, in 1965 the grease was made opaque with a mineral filler, to prevent photo electric effects and enhance heat conduction. Interestingly, germanium photo sensitivity peaks at longer wavelengths than silicon, 1500nm as opposed to 900nm. Fig.24 shows an OC42 with its metal sleeve removed. Note that most semiconductors are *increasingly unlikely* to fail the longer they have been in service; so, if a germanium device is still working, don't replace it just because it is old. Fig.25 shows an early seventies fuzz box using old STC copper-cased transistors which still worked, the rusty Omeg pots were long-gone however.

Fig.25. A 1974 Colorsound fuzz box using surplus copper-cased STC germanium transistors. Despite being corroded, they still worked fine

That's all we have space for this month – more on *Ge-mania* in the next issue!

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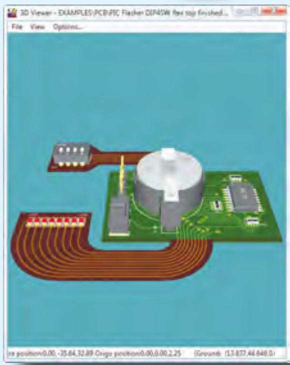

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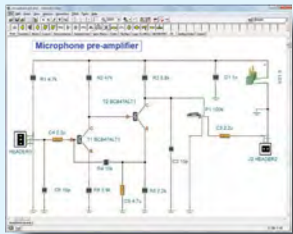

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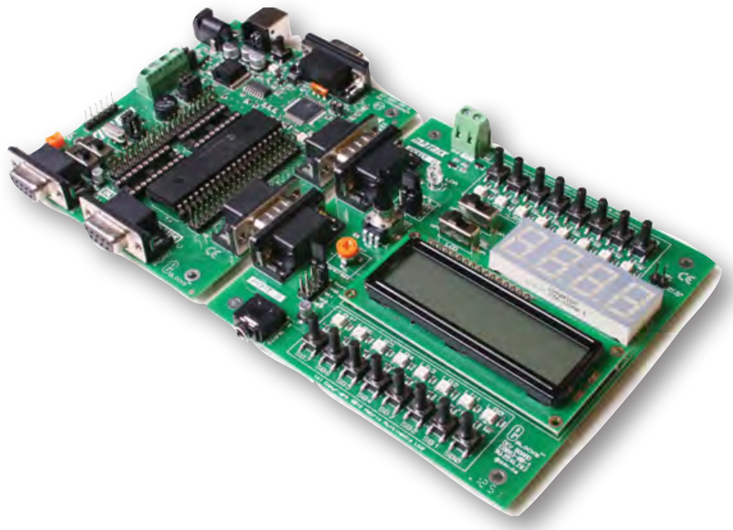

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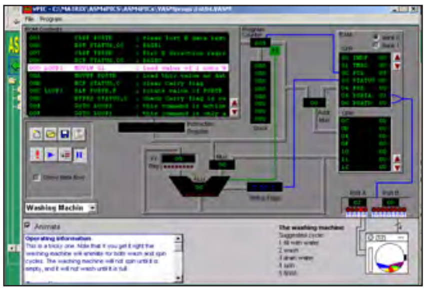

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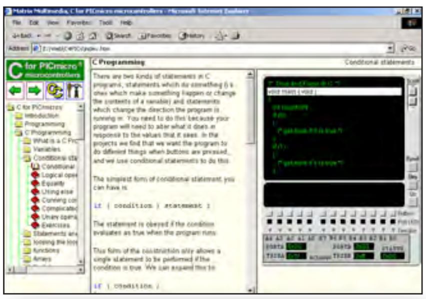

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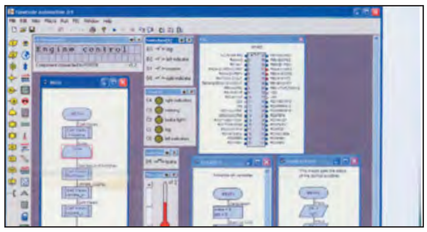

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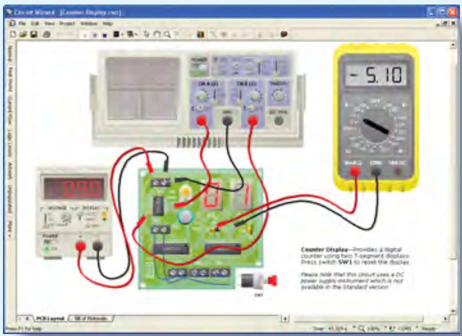

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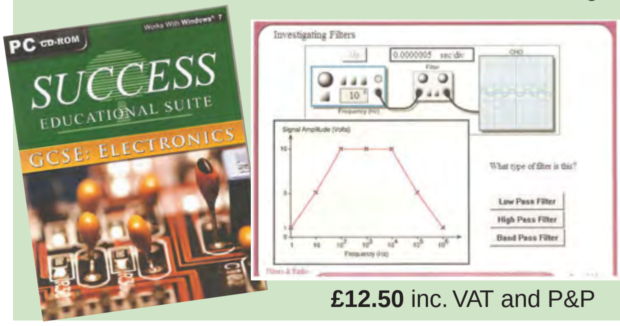

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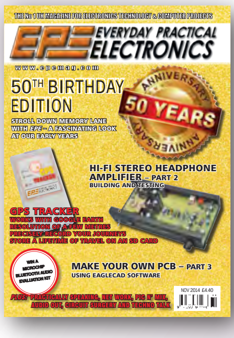

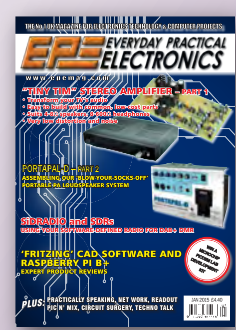

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Max's Cool Beans

By Max The Magnificent

Mastering meters – Part 4

As you will recall, we took a slight detour last month to take a sneak-peek at the forthcoming USB Type-C, but now it's time to return to the topic of using awesome antique analogue meters in one's hobby projects.

In Part 3 (see *EPE* March 2015) of this mini-series, we considered the circuit I use to drive my meters. This set-up includes a field-effect transistor to protect the output pin from the microcontroller (MCU) from back-electromotive force (EMF) effects and also – if necessary – to supply more current than the MCU can do on its own. The circuit also includes an external series resistor that we called R_{series}.

R_{series} reminder

In order to determine the R_{series} value for each meter, I created Max's super-duper *Series Resistance Test Unit*. This is really just a box with four linear potentiometers connected in series. If I were to make a new version of this box, the pot values I would use would be 500kΩ, 100kΩ, 10kΩ and 1kΩ. To be honest, I might also add a fifth 100Ω pot, because some meters require only a small R_{series} value (I have one that requires only 50Ω).

So we set all of these pots to their maximum values, then we use the MCU to drive its maximum pulse-width modulated value onto that output pin (this value is 255 in the case of my Arduino Mega). Now we slowly wind the pots down, starting with the largest value and working our way down to the smallest, until we get the meter's needle to reach its full-scale deflection (FSD). Then all we have to do is use a multimeter to determine the final resistance value of the test unit, and this will be our R_{series} value.

Now, when it comes to actually building the circuit on my Arduino's prototyping board, I implement the R_{series} value using a fixed 5% resistor in series with a multi-turn trim-pot (I try for 25 turns if possible). Both of these have a value as close as possible to R_{series}. I start with the trim-pot at its maximum value, drive a maximum PWM value out of the output pin, and gradually reduce the trim-pot until the meter reaches its FSD.

In Fig.1, we see a view of the insides of my Vetinari Clock, which is currently under construction (here's a video on YouTube: **http://bit.ly/1BUP50c**). My Arduino Mega is at the bottom. Next we have a large prototyping board carrying the circuits to drive my four meters (hence the four trim-pots; the fixed resistors are located behind the trim-pots). The smaller prototyping board on the top carries the real-time clock (RTC) and a temperature sensor. In the fullness of time, there will also be a sound effects card, but we will return to discuss that in a future column.

Actually, I perform an additional step before I implement the prototyping board. Once I've used the test unit to achieve FSD, I set the MCU to slowly cycle the output pin, driving the meter back and forth between the minimum and maximum PWM values (here's a video on YouTube: **http://bit.ly/1AMnJ5u**). If you watch your meter closely, you can spot if the needle sticks at all, in which case you may decide to set the meter aside for further evaluation.

Originally, I used a program that read the analogue value from an external potentiometer that I controlled by hand, and then used this value to modify the PWM output, but it's easy to miss tiny 'glitches' in the meter's movement using this technique. Getting the MCU to cycle things is much more efficacious.

About face!

Now, the thing about using antique analogue meters is that the chances of their faceplates reflecting the type of information you actually require is close to zero. In the case of my Vetinari Clock, for example, I wanted the faceplates to reflect units of time – 'Hours', 'Minutes' and 'Seconds' – as illustrated in Fig.2.

Another big consideration is the graphics themselves. There are all sorts of things you can do here. The cheapest and easiest is to simply create your own faceplate graphics using some sort of graphics program, print out these new graphics on good quality paper, cut them out, and stick them to the old faceplates, but I prefer to step things up for my projects.

First of all, I asked my graphics guru chum Denis Crowder (**http://bit.ly/1KucSZ2**) who hangs out in Hawaii, to create the graphic files

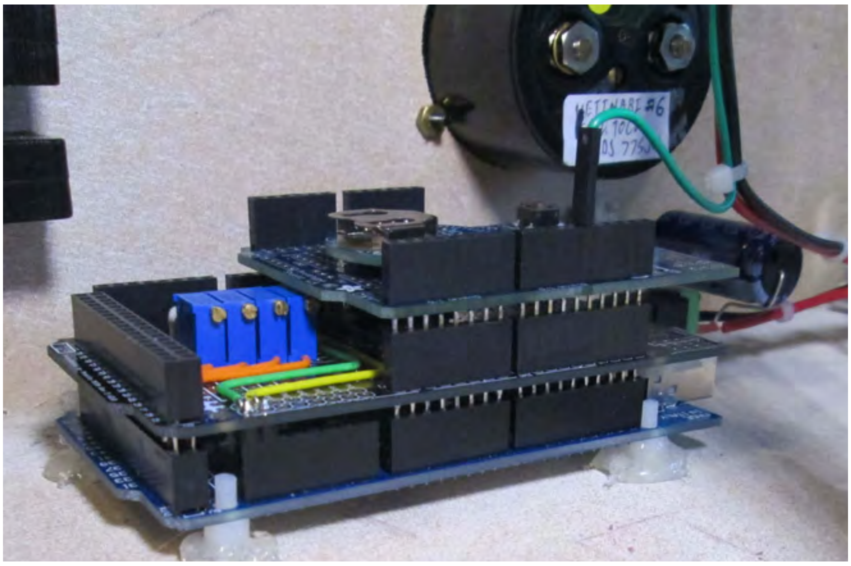

Fig. 1. The trim-pots for the four meters in my Vetinari Clock

themselves – my specific request was for an Art Deco look-and-feel. Next, master machinist John Strupat (**http://bit.ly/1wK80sI**), who hails from Canada, created new aluminium faceplates and printed the graphics onto them using a process he developed himself. The result looks like ceramic and is incredibly tasty. (John offers this as a service if you are interested in doing something similar yourself.)

Sad to relate, this is where I ran into a slight problem, because these antique analog meters are sealed units, and every time you open them up you run the risk of dust getting in. Even worse, when you start unscrewing things (like faceplates), if you aren't very careful, you can end up with metal particles sticking to the meter's magnet, which is not a good thing if they impede the meter's movement.

Fig. 2. The Vetinari Clock prototype jig; here we're testing the 'Hours' meter

But turn that frown upside down into a smile, because my chum Jason Dueck works for a company called Instrument Meter Specialties (IMS: **http://bit.ly/1B8MAVM**), which is based in California. These folks are expert at refurbishing antique meters for use in aircraft, submarines, nuclear power stations, industrial plants, and so forth. Thus, I ended up sending all of my meters to the folks at IMS, and they stripped them down, gave them a good servicing, replaced the original faceplates with the spiffy new versions from John, and returned the little beauties to me.

There's so much more to talk about here, but we'll save all that as a treat for a future column. In the meantime, I will be presenting a paper on all of this at the Embedded Systems Conference (ESC) in Boston in May, and I just created a short video that talks about this and also shows some other meter-based projects (**http://bit.ly/1Ma2voP**). *Until next time, have a good one!*

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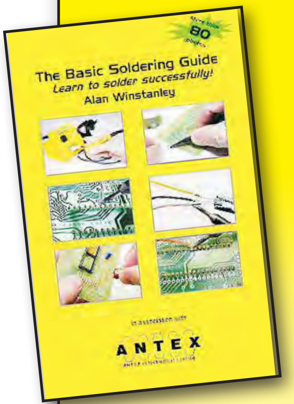

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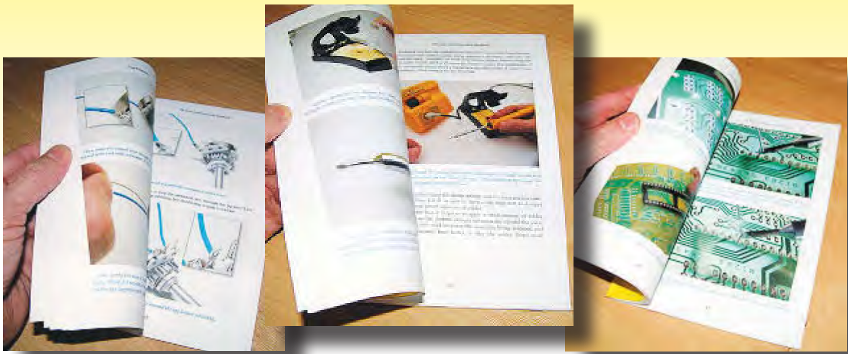

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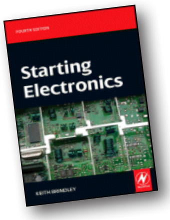

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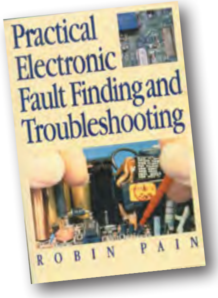

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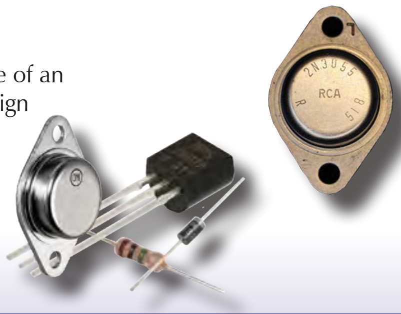
50 Years of the 2N3055

The 2N3055 NPN power transistor appeared in an RCA publication in 1964, so in 2014, it was 50 years old! Few other transistors have survived as long, and it is perhaps ironic that RCA are no longer around to celebrate. John Ellis discusses this iconic device.

Teach-In 2015 – Part 5

In June's *Teach-In 2015*, we introduce the use of filters to modify the frequency response of an amplifier; take a detailed look at noise and how it can be reduced; and describe the design and construction of a versatile tone control stage that can be used on its own or in conjunction with the other projects featured in *Teach-In 2015*.



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